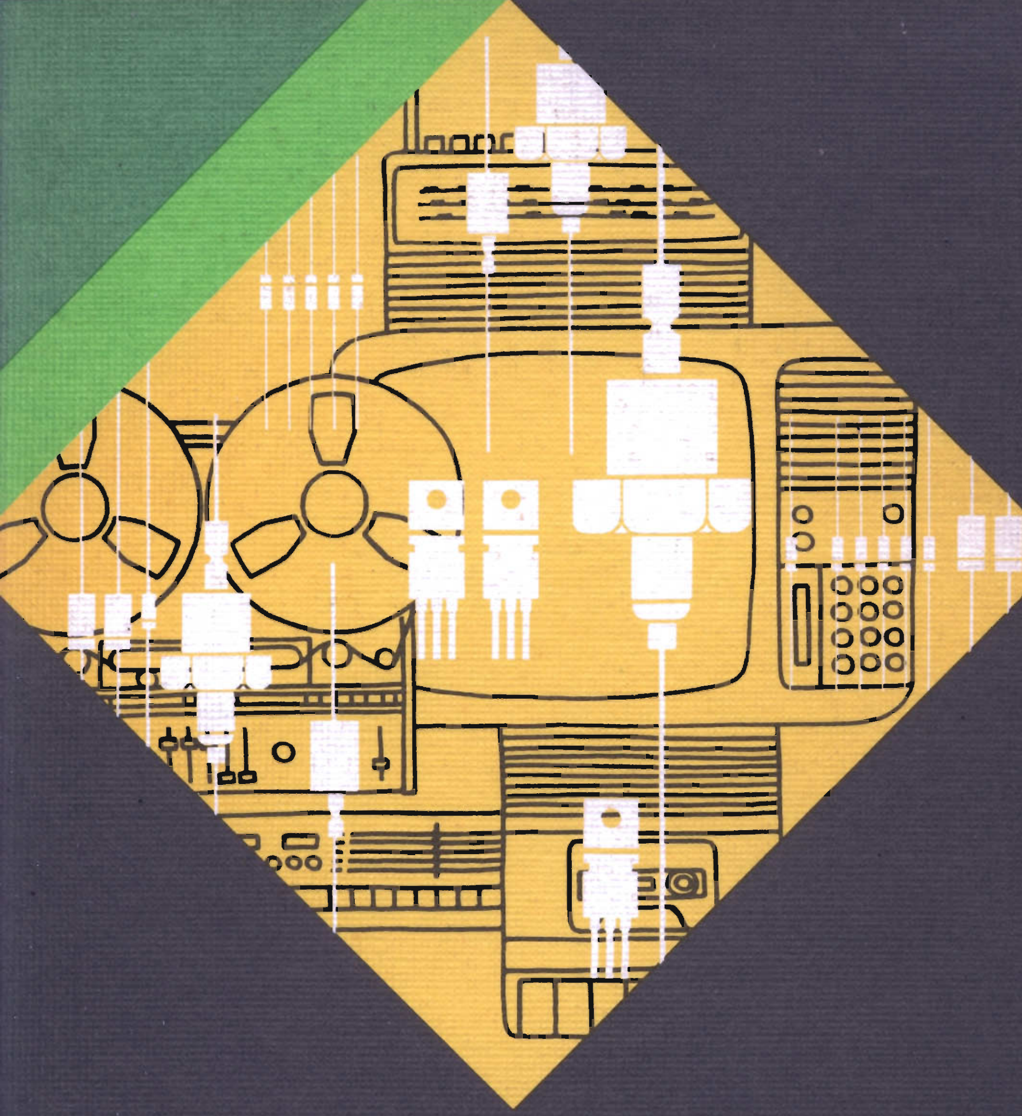


Zener Diodes, Integrated Stabilising Circuits and Voltage Regulators Basics and Applications

6220-11-1E

semiconductors ITT

ITT

Zener Diodes, Integrated Stabilising Circuits and Voltage Regulators

Basics and Applications

compiled by
Ing. (grad.) Rudolf Sydow
from publications by the following
members of INTERMETALL
Dr.-Ing. Albrecht Gerlach
Dr. rer. nat. Ulrich Dolega
Dipl.-Ing. Wolfgang Höhn
and others

Edition 1978

© 1978 International Telephone
and Telegraph Corporation

Reprinting is generally permitted, indicating the source. However, our consent must be obtained in all cases. Information furnished by ITT is believed to be accurate and reliable. However, no responsibility is assumed by ITT for its use; nor for any infringements of patents or other rights of third parties which may result from its use. No license is granted by implication or otherwise under any patent or patent rights of ITT. The information and suggestions are given without obligation and cannot give rise to any liability; they do not indicate the availability of the components mentioned. Delivery of development samples does not imply any obligation of ITT to supply larger amounts of such units to a fixed term. To this effect, only written confirmation of orders will be binding.

Printed in W.-Germany · Imprimé dans la République Fédérale d'Allemagne
by Druckhaus Rombach+Co GmbH, 7800 Freiburg

Edition 1978/2 · Order No. 6220-11-1 E · Subject to modifications

Preface

As manufacturers of Zener diodes [1] INTERMETALL boasts a tradition going back 20 years [2], [3]. Taking the first Zeners developed in 1956 (types Z 6, Z 7 and Z 8) as a point of departure, a broadly based program of Zener diodes for every purpose has grown over the years, supplemented and rounded off by the temperature-compensated Zener diode and by monolithic integrated stabilising circuits.

Without laying claim to completeness, this booklet aims at familiarising the practising engineer and the student with Zener diodes and stabilising circuits, offering dimensioning guidelines and suggestions for the solution of stabilising problems by way of circuit examples.

Data relating to INTERMETALL semiconductors discussed here and contained in the circuit examples will be found in the INTERMETALL Semiconductor Summary 1978/79. The supply of INTERMETALL semiconductors in small and medium quantities is the responsibility of INTERMETALL distributors. Please address enquiries concerning prices and deliveries to your local distributor.

Contents

		Page
	Preface	5
1.	**Introduction**	8
1.1.	General	8
1.2.	Passive stabilising components	8
1.2.1.	Inductive voltage stabilisers	8
1.2.2.	Neon stabilisers	9
1.2.3.	Corona stabilisers	10
1.2.4.	PTC resistors	10
1.2.5.	NTC resistors	10
1.2.6.	Zener diodes	10
1.2.7.	Silicon diodes or rectifiers, light emitting diodes	11
1.2.8.	Temperature-compensated Zener diodes, stabiliser diodes	12
1.3.	Active stabilising or voltage regulator circuits	12
2.	**Zener diodes and similar components**	14
2.1.	**Zener diodes**	14
2.1.1.	Physical principles	14
2.1.2.	Manufacturing processes	16
2.1.3.	Electrical properties	19
2.1.3.1.	Operating characteristics, temperature coefficient	19
2.1.3.2.	Thermal resistance	24
2.1.3.3.	Differential resistance	24
2.1.3.4.	Capacitance	26
2.1.3.5.	Noise	27
2.1.4.	Measuring the operating voltage of Zener diodes	28
2.1.5.	Design of stabiliser circuits with Zener diodes	28
2.1.5.1.	Stabilisation factor, source resistance	28
2.1.5.2.	Operating voltage, operating characteristic	30
2.1.5.3.	Power dissipation, thermal resistance, cooling measures	36
2.2.	**Stabiliser diodes**	40
2.2.1.	The structure of stabiliser diodes	40
2.2.2.	Data of stabiliser diodes ZTE 1,5 . . . ZTE 5,1	41
2.3.	**Temperature-compensated Zener diodes**	44
2.3.1.	Methods of temperature compensation	44
2.3.2.	The construction of ZTK diodes	45
2.3.3.	Data of the ZTK diodes ZTK 6,8 . . . ZTK 33	48
3.	**Monolithic integrated voltage regulators**	52
3.1.	General	52
3.2.	Voltage regulators of the series TDD 1605 . . . TDD 1624	52
3.2.1.	Construction	52
3.2.2.	Data for voltage regulators TDD 1605 . . . TDD 1624	54
3.3.	The car voltage regulator TCA 700 X	56
3.3.1.	Demands placed upon this device	56
3.3.2.	The TCA 700 X – Design considerations	56
3.3.3.	Technical data of the TCA 700 X	57
3.4.	Cooling of integrated voltage regulators	58

		Page
4.	**Application circuits**	60
4.1.	Circuit arrangements with Zener diodes, ZTE and ZTK diodes	60
4.1.1.	Parallel stabiliser circuits	60
4.1.2.	Bridge stabilisers	61
4.1.3.	Voltage stabiliser circuits with Zener diode and transistor	62
4.1.4.	Current stabilising circuits with Zener diode and transistor	64
4.1.5.	Alternating current circuits with Zener diodes	66
4.1.6.	Limiting and protective circuits	69
4.1.7.	Various circuit arrangements featuring Zener diodes	72
4.2.	Voltage regulator circuits	75
4.2.1.	Basic voltage regulator circuits	76
4.2.2.	Overload protection for voltage regulator circuits	81
4.2.3.	Circuit examples for voltage regulators	83
4.2.4.	Circuit examples for voltage regulators with integrated circuits	88
4.3.	Laboratory power supply unit with voltage and current regulation	90
5.	**Bibliography**	92

1. Introduction

1.1. General

The need to keep the supply voltage for critical parts of a plant or a piece of equipment constant, irrespective of battery or mains voltage fluctuations and, usually, also irrespective of load variations has been one of the early requirements of electrical engineering. Two basically different methods of solving the problem were adopted. There is the direct stabilisation of voltage or current by passive components having a non-linear characteristic. Alternatively, where the demands and the required performance are higher, direct stabilisation of voltage or current is chosen, using active correcting elements which are actuated by a control amplifier, the latter comprising again a component with non-linear characteristic as a reference voltage source.

1.2. Passive stabilising components

The following components may serve as voltage or current stabilisers because of the non-linearity of their *V/I* characteristic.

1.2.1. Inductive voltage stabilisers

In the case of inductive voltage stabilisers [4], the non-linear magnetising characteristic (Fig. 1) of the iron core of saturable reactors is exploited

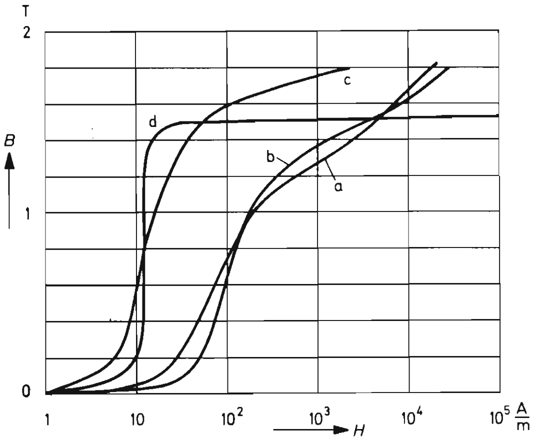


Fig. 1: Magnetisation characteristics for transformer sheets [5]
a) Dynamo sheet DIN 46 400 III
b) Transformer sheet DIN 46 400 IV/1.0
c) Oriented silicon iron
d) Material with square magnetisation characteristic (F3) according to DIN 41 301 (50 % Ni)

for stabilising purposes. These devices are suitable for stabilising the AC mains voltage for low-power loads.

1.2.2. Neon stabilisers

Neon stabilisers [4] are gas discharge tubes with a current/voltage characteristic as shown in Fig. 2, frequently used in thermionic valve equipment for the stabilisation of DC supply voltages, either in direct (passive)

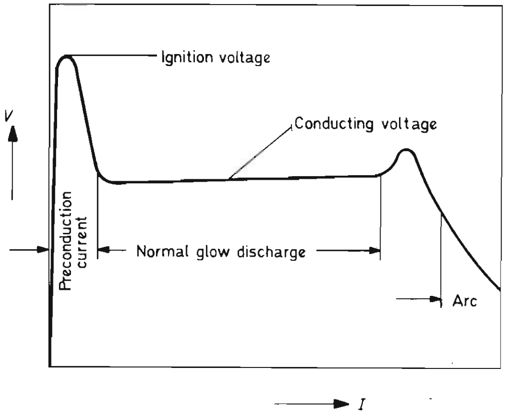


Fig. 2: Characteristic of a gas discharge path (schematic)

parallel stabilising circuits (Fig. 3) or in (active) control circuitry – in the latter case serving as reference voltage source [6].

The conducting voltage of such a gas discharge path is between 70 and 150 V, the ignition voltage amounts to approximately 1.5 times this value. When the equipment is switched on, the supply voltage $+V_B$ of Fig. 3 should be at least 50 % higher than the conducting voltage of the stabiliser, to enable the latter to ignite. Frequently, multi-path stabilisers, e. g. STV 280/40, were employed which, in one and the same glass envelope, contain four series-connected discharge paths. Thus, stabilised supply voltages of 70 V, 140 V, 210 V and 280 V were obtained for the equipment in question. Parallel stabilising circuits – see Fig. 3 – have a comparatively low efficiency and are not suitable for stabilised voltages lower than about 70 V. An advantage is the low temperature coefficient of the conducting voltage – approximately $3 \times 10^{-5}/°C$.

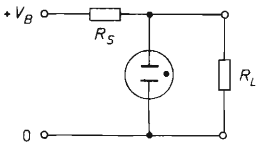


Fig. 3:
Parallel stabilising circuit with neon stabiliser

1.2.3. Corona stabilisers

Corona stabilisers [7] resemble neon stabilisers as regards characteristic and operation, but their conducting voltages amount to several hundred volts. They operate with currents of only about 50 μA and are employed for the stabilisation of counter tube supply voltages.

1.2.4. PTC resistors

"PTC resistor" [4] is a generic term for resistors with a (large) positive temperature coefficient. There are T-PTC resistors, comprising a resistance wire (iron or tungsten), fused into an evacuated or gas-filled glass envelope, and S-PTC resistors consisting of a semiconductor material, the latter being available in two variants. Based on silicon, PTC resistors are being offered which have a comparatively small positive temperature coefficient extending over a wide range of temperatures [8]. Based on barium metatitanate, PTC resistors are being manufactured which have a very large positive temperature coefficient that covers only a narrow range of temperatures, however.

Iron-hydrogen resistors were of practical importance in the thirties as biasing resistors stabilising the filament current in universal mains receivers. Today, they are still used on rare occasions, partly as commercial incandescent lamps, partly as professional PTC resistors in electronic stabilising circuits, e. g. for amplitude stabilisation in *RC* generators [9].

1.2.5. NTC resistors

NTC resistors, also referred to as thermistors [8] are resistors having a large negative temperature coefficient. They are produced from sintered semiconducting oxides, Although they have a non-linear V/I characteristic in the static condition [8], they are scarcely used for stabilising purposes.

1.2.6. Zener diodes

Zener diodes have a current/voltage characteristic such as indicated in Fig. 4 [10]. In its left part, this resembles the characteristic of a neon stabiliser (Fig. 2), the advantage being, however, that the Zener diode does not require an ignition voltage, so that in a stabilising circuit (Fig. 5) corresponding to the one shown in Fig. 3, the difference between the supply voltage V_B and the stabilised voltage may be smaller than in the case of a neon stabiliser circuit. Zener diodes are manufactured with operating voltages from about 3 V to 200 V. Apart from direct passive stabilising circuits such as shown in Fig. 5, Zener diodes are frequently employed as reference voltage sources in active control arrangements.

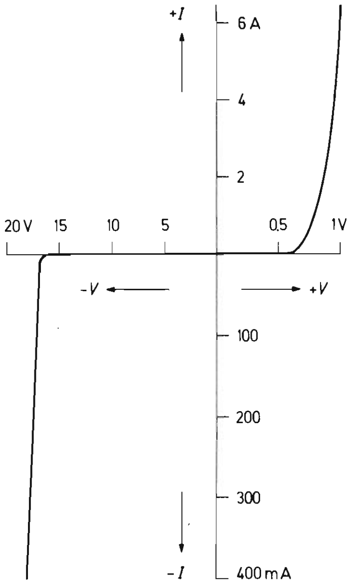


Fig. 4:
Characteristic of a
silicon power Zener diode

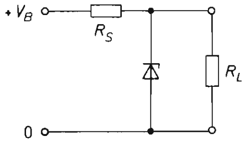


Fig. 5:
Parallel stabilising circuit
with Zener diode

1.2.7. Silicon diodes or rectifiers, light emitting diodes

Silicon diodes or rectifiers are likewise characterised by a non-linear characteristic [10], such as shown in Fig. 6. Here, the forward characteristic (Fig. 6, right) ist used to generate, as shown in Fig. 7, stabilised voltages of about 0.7 V (one diode), 1.4 V (two diodes in series) or 2.1 V (three diodes in series). For this voltage range, no useful Zener diodes are available for technological reasons.

Because of their characteristic, light emitting diodes [23] are suitable for the circuit shown in Fig. 7. It is an interesting phenomenon that their forward voltage exhibits the same absolute temperature dependence as that of silicon diodes. Since the forward voltage of light emitting diodes is two to three times as high as that of silicon diodes, their relative temperature dependence (their temperature coefficient) amounts only to half or one third of the temperature coefficient of the forward voltage in normal silicon diodes.

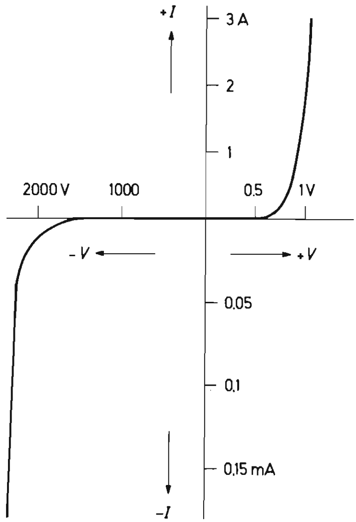


Fig. 6:
Characteristic
of a 1 A silicon rectifier

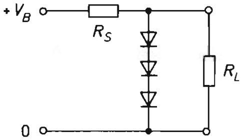


Fig. 7:
Parallel stabilising circuit
for 2.1 V with three series-connected
silicon diodes

1.2.8. Temperature-compensated Zener diodes, stabiliser diodes

Temperature-compensated Zener diodes type ZTK [11] and stabiliser diodes, type ZTE are "improved" Zener diodes, developed by INTERMETALL, for certain voltages and special uses, which are employed in circuits like that of Fig. 5. They are integrated circuits with two terminals which, electrically, behave as Zener diodes, but have a lower temperature coefficient and a lower dynamic resistance.

1.3. Active stabilising or voltage regulator circuits

In many cases, given demands cannot be met by passive stabilising circuits (Figs. 3, 5 or 7), because the stabilising effect is too weak, dissipation too high or efficiency too low. In that case, circuit arrangements such as that shown in Fig. 8 are indicated. Here, the output voltage of the circuit is maintained at the desired level by the final control element. The difference between nominal and actual values is boosted by a control amplifier whose output signal influences the final control element in the required sense, until the actual value coincides with the nominal

value. Possible final control elements are toroidal variable-ratio transformers, transductors [12], vacuum tubes [6], transistors and possibly also thyristors or triacs [13]. This book deals only with direct voltage and direct current control circuits, using a transistor as the final control element. Switch-mode controls are not discussed.

Nearly all previously discussed passive stabilisers are usable as a reference voltage source in Fig. 8, Zener diodes having proved the most popular stabiliser. For low to medium power applications, integrated regulators are usually employed today [14] and represent an extremely economical solution for electronic voltage regulators.

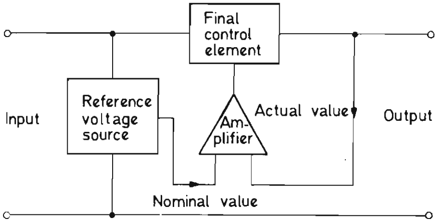


Fig. 8: Block diagram of an (active) voltage regulator circuit

2. Zener diodes and similar components

2.1. Zener diodes [15]

2.1.1. Physical principles

The reverse characteristics of all semiconductor diodes exhibit similarity of behaviour. The reverse current at first rises only insignificantly with the rising reverse voltage. It is only when a certain threshold is exceeded that it rapidly attains high values, especially in silicon diodes. Investigations by Dr. C. Zener [16], undertaken 40 years ago on breakdown phenomena in a solid dielectric have contributed to the elucidation of the physical events in this breakdown region. However, the Zener effect does not determine the characteristic of breakdown voltage above about 5.5. V, the determining factor under these conditions being, inter alia, the avalanche breakdown effect, according to a theory put forward by McKay [17].

Fig. 9 is a diagrammatic illustration of the distribution of impurities and charge carriers in a PN junction to which a reverse voltage has been applied. The silicon atoms of the monocrystal are not shown in this drawing. In the N region, the solid plus signs symbolise the donors, e. g. 5-valent antimony atoms, while in the P region the acceptors included in the silicon lattice – for example, 3-valent aluminium atoms – are represented by solid minus signs. At room temperature, both kinds of impurity atoms are almost completely ionised, i. e. the amount of freely moving electrons (thin minus symbols) in the N region is roughly the same as that of donors. The same applies to the amount of holes (thin plus symbols) in the P region. When influenced by a reverse voltage such as that applied according to Fig. 9a, the holes move in the direction towards the negative terminal of this voltage and the electrons towards the positive terminal. In this way, a zone free from moving charge carriers is created near the border between the P and the N region, and the ionised impurity atoms fixed in the silicon lattice form a space charge.

In this simplified model, the fact that it is not possible to produce a PN-junction as abrupt as shown in the drawing has been neglected. Nor is the border of the region vacated by the charge carriers sharply defined: Due to the fact that the free charge carriers execute a thermal diffusion movement this limit is rather diffuse. Furthermore, the concentration of donors is different from that of acceptors in the case of normal diodes. Finally, a small number of charge carriers is supplied by the silicon atoms due to thermal ionisation. These electron/hole pairs generate a small reverse current even with a low reverse voltage.

Notwithstanding these neglected facts, the electrical phenomena which take place in the depletion layer can be explained with reference to the drawing of this simplified model.

Plotting the space charge density q, generated by the ionised impurity atoms, on the positional co-ordinate x, the square wave distribution shown in Fig. 9b is obtained. At larger distances from the PN junction, an equilibrium has been established between the ionised impurity atoms

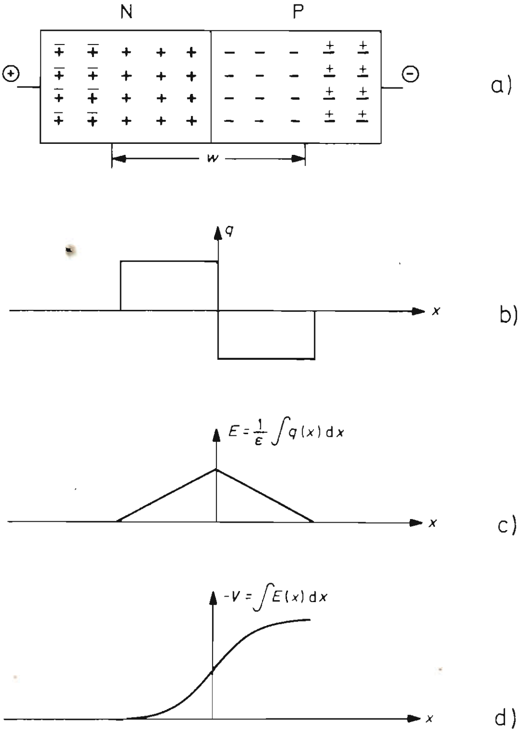


Fig. 9: Reverse biased PN junction
a) Space charge distribution
b) Space charge density plotted on the positional co-ordinate
c) Field strength at the depletion layer, plotted on the positional co-ordinate
d) Voltage distribution on the positional co-ordinate.

and the freely moving charge carriers. There is no space charge, in that case. The resulting field strength may be calculated by integration via the space charge density (Fig. 9c), and the potential distribution by repeating the integration (Fig. 9d).

$$E_{max} = \frac{qw}{2\varepsilon} \qquad (1)$$

$$V_r = \frac{qw^2}{4\varepsilon} \qquad (2)$$

Equation 1 indicates the maximum field strength at the depletion layer. w is the width of the region depleted of charge carriers, and ε is a constant. According to equation 2, the inverse voltage between the limits of

this zone can be calculated. If w is eliminated from both equations and, moreover, the charge carrier density q is replaced by the specific resistivity ϱ of the semiconductor material, a far more customary value which is equal to the reciprocal of the product from charge carrier concentration (space charge density) and carrier mobility, then we obtain for the maximum field strength

$$E_{max} = \sqrt{\frac{V_r}{\varepsilon \mu \varrho}} \tag{3}$$

It will be seen that the breakdown field strength which, in the case of silicon, amounts to approximately 500 kV/cm is reached as a result of increasing the reverse voltage the sooner the more low-ohmic the semiconductor material used in the manufacture of the Zener diode. Equation 2 reveals that the width of the charge carrier depleted zone grows with the reverse voltage. This effect is exploited in the case of variable capacitance diodes [18].

Because of the neglected factors, equation 3 does not reveal the relationship between the resistivity of the basic material and the breakdown voltage accurately; it suggests only the trend. In the table 1, given below, the resistivities of the basic material and the width of the depleted zone on breakdown are quoted for several breakdown voltages.

Table 1:
Relationship between the operating voltage, the resistivity of the basic material and the width of the depleted zone on breakdown.

Operating voltage V_Z V	Resistivity ϱ Ωcm	Width of the depleted zone w µm
3.5	0.01	0.1
12	0.1	0.6
50	1	4

The varying width of the depletion layer is one of the reasons why the thermally generated reverse current increases a little as the reverse voltage increases in the region below the level of the operating voltage. Besides, surface contamination contributes to the rise of the reverse characteristic.

2.1.2. Manufacturing processes

All electrical phenomena connected with the events taking place in the semiconductor crystal can be influenced within wide limits by varying the impurity atom concentration (donors or acceptors) on either side of the PN junction, by varying the thickness of the P-type conductivity and N type conductivity regions, by varying the crystal area and by selecting the manufacturing process. Manufacturing processes differ in the way the desired donor and acceptor concentrations are introduced in the basic silicon material [10].

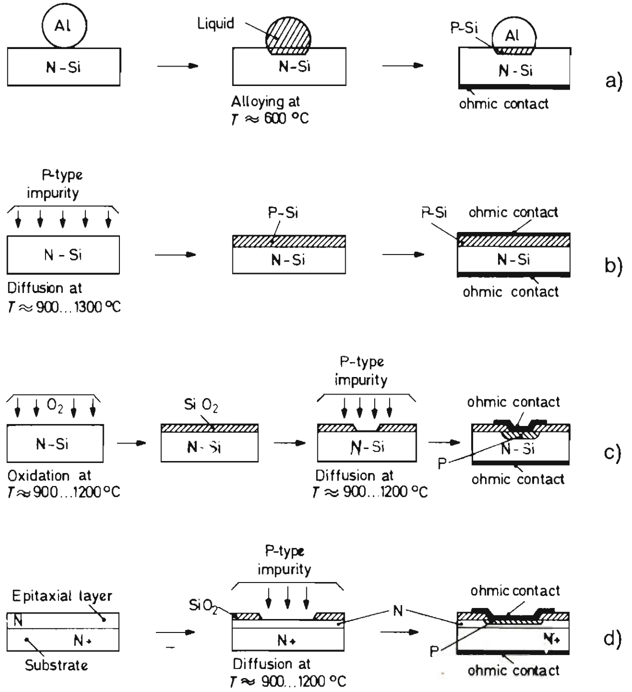


Fig. 10: Manufacturing processes for PN junctions
a) Alloying process
b) Diffusion process
c) Planar diffusion process
d) Epitaxial planar diffusion process

In the alloying process (Fig. 10a) a very small aluminium ball or a piece of aluminium wire is placed upon an N doped silicon wafer. Both these elements are heated in a furnace to a point slightly above the eutectic temperature, so that a small region of molten AlSi alloy is produced. On cooling, this alloy solidifies, producing an aluminium-saturated, re-crystallised P type conductivity zone. The remaining aluminium may be used as the ohmic P type zone contact. On the back, an ohmic contact may be obtained by vapour depositing an antimony doped gold layer, which is alloyed at a temperature of approximately 400 °C, the antimony (Sb) producing a Si zone of pronounced N type conductivity. The exact positions, the area and the depth of penetration of a PN junction produced in this way depend very much on the temperature/time cycle during the alloying process and are difficult to control accurately. This is why the alloy technology – originally the only process available – is today used only in special cases, e.g. in the manufacture of power Zener diodes with operating voltages below 10 V. Elsewhere it has been superseded by improved processes.

The solid-state diffusion process, developed later than the alloying process, allows a much better control of the impurity atom concentration profile and thus of the penetration depth of the PN junction (see Fig. 10b). Monocrystalline N type Si wafers of 5 to 7.5 cm diameter are exposed at high temperatures (900 . . . 1300 °C) to an atmosphere containing a doping agent (e. g. boron in the shape of BBr_3). The impurity atoms are diffused into the N type silicon, and the penetration depth (1 . . . 50 μm) can be controlled very accurately by a careful choice of temperature and time. However, the area of the PN junction used for the individual diode cannot be determined by this process. Instead, the Si wafers, after being provided with ohmic contacts by galvanic or non-galvanic metallisation or by vapour deposition of suitable metal layers, are divided into chips of the desired size either by scribing and breaking or by sawing. In that case, the laterally exposed PN junction must still be etched and covered. This process is now employed for power Zener diodes with operating voltages above 10 V.

The control of the geometry of PN junctions is improved by several orders of magnitude in the so-called planar process (1960). A typical example is shown in Fig. 10c. On the surface of the N type silicon wafer, a SiO_2 layer, about 1 μm thick, is produced in oxygen at high temperatures. This layer acts as a mask for the doping agents. By means of a photolitho process, windows are produced in this SiO_2 layer through which impurity atoms are diffused into the N type silicon during the subsequent diffusion step. With this method, PN junctions with as little as 3 μm side length can be produced. Instead of the SiO_2 layer, sometimes an Si_3N_4 layer, or a combination of both types of layer, is used at present.

A further improvement over the planar diffusion method as regards the series resistance encountered in the original N type silicon layer (thickness: 100 . . . 200 μm) is achieved in the epitaxial process (from the Greek EPI = upon and TAXIS = arrangement). In this process, a usually high-ohmic monocrystalline Si layer is deposited in a chemical reaction from the gaseous phase upon a low-ohmic, likewise monocrystalline silicon wafer substrate. The further steps of this process are the same as in the

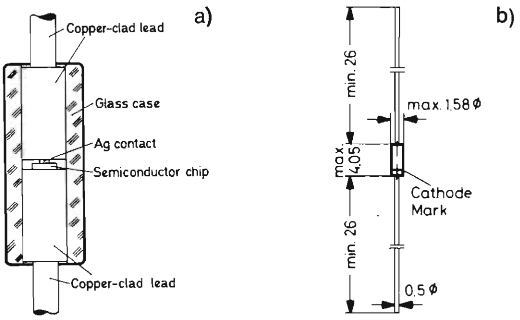


Fig. 11: Zener diode in the DO-35 glass case
a) Sectional view b) Dimensional diagram

planar diffusion process (see Fig. 10d). Today, 0.5 and 1 W Zener diodes are manufactured by INTERMETALL in planar or epitaxial planar technology for all voltages.

Fig. 11 shows the construction of a modern 0.5 W Zener diode in the DO-35 glass case (54 A 2 according to DIN 41 880). With this case, the U oder S bends, needed for the earlier DO-7 case, are superfluous and the required contact pressure is produced after fusing by the glass case when it shrinks during cooling. INTERMETALL power Zener diodes of the series ZX, with a maximum dissipation of 10 W, are supplied in a metal case with a threaded stud which enables screw-connection of the diode to a cooling plate or a heat sink. Fig. 12 shows the construction of such a diode.

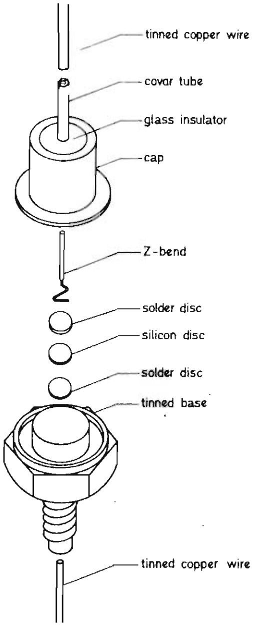


Fig. 12: Construction of a power Zener diode (series ZX . .)

2.1.3. Electrical properties

2.1.3.1. Operating characteristic, temperature coefficient

Since, normally, the forward characteristic of Zener diodes is of no interest, usually only the breakdown or the operating characteristic is indicated in data sheets and books. This is the part of the characteristic, shown in Fig. 4 in the third quadrant of the co-ordinate system. To simplify the presentation, the drawing of Fig. 4 is turned through 180°, so that the operating characteristic comes to be situated in the first quadrant, see Fig. 13. This drawing shows diagrammatically the operating

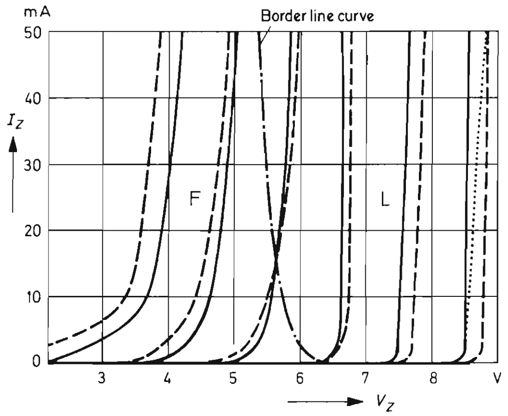


Fig. 13: Operating characteristics of various Zener diodes in the 4 . . . 9 V range, measured by means of pulses at a constant junction temperature
——— T_j = 25 °C
- - - - T_j = 100 °C
. . . . . plotted under static conditions

characteristics of some Zener diodes at junction temperatures of T_j = 25 °C (solid lines) and T_j = 100 °C (dash-lines).

For a proper understanding it is important to note that these characteristics are valid for a constant junction temperature. In order to plot them, one of two possibilities may be adopted. Either the Zener diode is fed with current pulses of short duration and the voltage drop which occurs each time across the diode is measured. The characteristic is plotted point by point, without the diode being heated. Alternatively, the characteristic is plotted by means of an electronic curve tracer in single sweep operation, in which case the current pulse must be sufficiently short if the junction temperature is not to vary. Plotting the operating characteristic with direct current produces a curve which, at low currents, is identical with the characteristic for T_j = 25 °C and which, at some current I_Z dependent upon the thermal resistance, coincides with the characteristic for T_j = 100 °C. This is indicated by a dotted line on the right in Fig. 13.

With Zener diodes, functioning with voltages below 6 V, the operating characteristic shifts to the left with rising temperature – the operating voltage has a negative temperature coefficient – and, in the case of Zener diodes functioning with operating voltages above 6 V the temperature coefficient of the operating voltage is positive. With an operating voltage of about 5.5 V, the temperature coefficient changes its sign, as is shown by the curves in Fig. 14. If the inherent differential resistance r_{zj}, which is a measure of the slope of the operating characteristic, is shown as a function of the operating voltage (Fig. 15), then this curve exhibits a pronounced minimum at about 7 V.

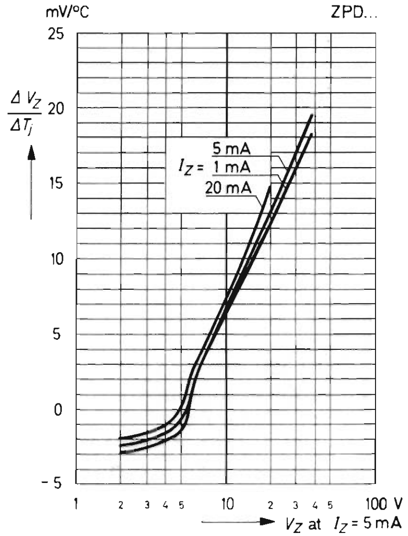


Fig. 14:
Temperature response of the operating voltage versus operating voltage

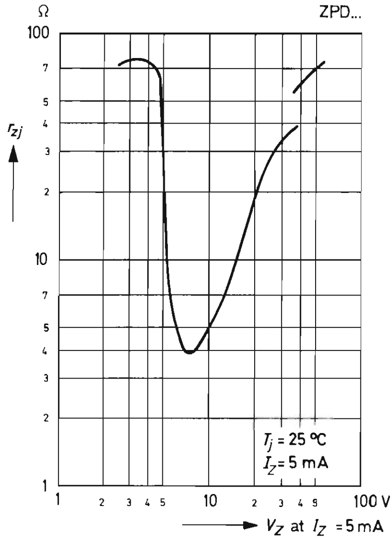


Fig. 15:
Inherent differential resistance versus operating voltage

The facts expressed in Figs. 14 and 15, as well as the far more pronounced bend, made apparent in Fig. 13, of the operating characteristic of Zener diodes with operating voltages above 6 V suggest that the physical effect responsible for producing the operating characteristic in Zener diodes with operating voltages below 6 V is different from the one responsible for producing the operating characteristic in Zener diodes functioning with operating voltages above 6 V.

With the so-called Zener effect [16], which dominates in Zener diodes below 6 V, field emission begins to occur as a critical field strength is exceeded. Valency electrons of the silicon atoms are raised to the level of the conductivity band, although their energy is less than the ionisation energy required according to classic observations. However, by resorting to the equations of quantum mechanics, a certain ionisation prohability can be computed. It is said that "tunnelling through the potential wall occurs". The required ionisation energy decreases with rising temperature. This is why the above mentioned prohability of the production of free charge carriers increases with temperature. Since it is only Zener diodes with operating voltages of less than 6 V that exhibit a negative temperature coefficient, it is assumed that the Zener effect plays an essential part only in this region.

The positive temperature coefficient of the operating voltage, which comes into play at higher levels of this voltage suggests that here the

so-called "avalanche breakdown" occurs. This phenomenon rests on the fact that thermally produced charge carriers, always present in the depletion layer, are accelerated by the available field strength to an extent sufficient for their kinetic energy to ionise silicon atoms. The new carriers thus produced are likewise accelerated and, in turn, ionise further atoms. A chain reaction occurs. The likelihood of this event taking place increases with the path length of the charge carriers, i. e. the distance a charge carrier travels on average without colliding with a unit cube. Owing to the increasing thermal movement of the unit cubes, it decreases with rising temperature. This produces the positive temperature coefficient of the operating voltage in the case of avalanche breakdown.

In Fig. 13, the region in which avalanche breakdown dominates (L) is separated from the region in which the Zener effect (F) operates, by a border line curve. If it is desired to provide a low cost reference voltage source without temperature dependence, a Zener diode with an operating voltage of about 5.7 V may be employed to advantage. Fig. 16 shows how operating voltage varies with temperature in such a diode and makes it clear that it is possible, by a suitable choice of the operating voltage and the operating current, to restrict the operating voltage to insignificant variations over a wide range of temperatures.

Fig. 17 shows the variation with junction temperature of the operating voltage of Zener diodes with different operating voltages. As this diagram shows, the temperature coefficient remains constant over the whole range of temperatures for Zener diodes functioning at voltages above 7 V, whereas in the case of Zener diodes functioning at lower voltages the temperature coefficient is dependent on temperature. The straight line for $V_Z = 1$ V belongs to a silicon diode of the ZPD 1 type, operating in the forward direction.

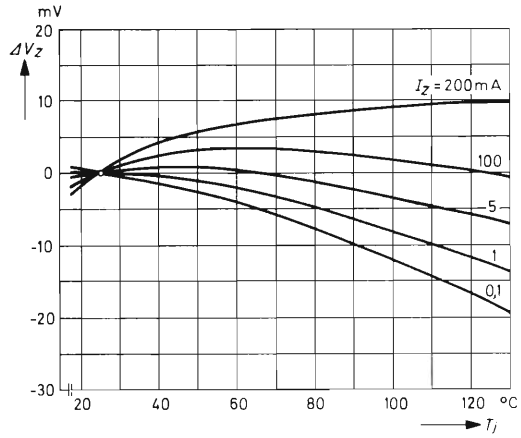


Fig. 16: Variation of the operating voltage with junction temperature in a 5.7 V Zener diode, related to $T_j = 25$ °C

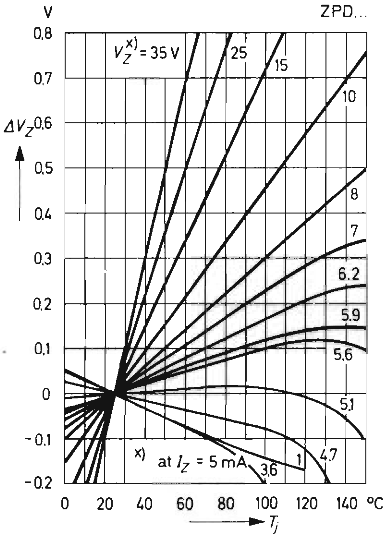


Fig. 17:
Variation of the operating voltage versus junction temperature

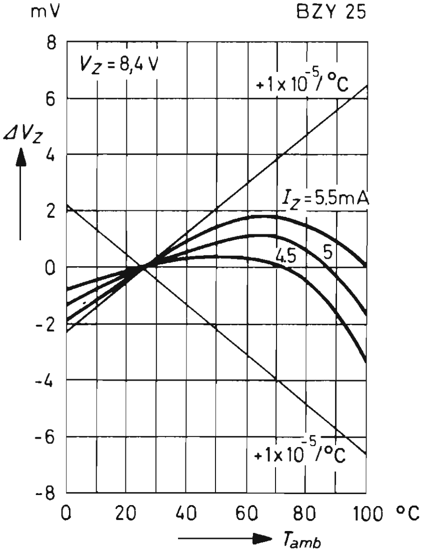


Fig. 19:
Variation of the operating voltage versus the ambient temperature in the case of a BZY 25 reference element

For applications in which the temperature dependence of the operating voltage is required to be very small, INTERMETALL produced so-called silicon reference elements [20] in the sixties in which the positive temperature coefficient of a 7 V Zener diode is compensated by two such silicon diodes operating in the forward direction were connected in series, and all three diodes were thermally coupled in a common metal case (Fig. 18). An operating voltage of 8.4 V and the temperature dependence of the operating voltage shown in Fig. 19 were the result. The INTERMETALL designers went still one step further by developing the reference amplifier, revolutionary at the time (1962), in which one of the forward diodes, provided for compensating purposes, was formed by the base emitter diode of a transistor (Fig. 20), so that for application in voltage regulator circuits a single component became available which contained in one and the same case the temperature-compensated reference

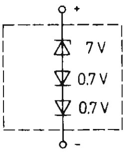


Fig. 18:
Internal circuit of a BZY 25 reference element

voltage source and the control amplifier. These reference elements and reference amplifiers were later superseded by temperature-compensated Zener diodes of the series ZTK [11], developed towards the end of the sixties by INTERMETALL, and by the integrated voltage stabilisers [21] which have meanwhile found wide application.

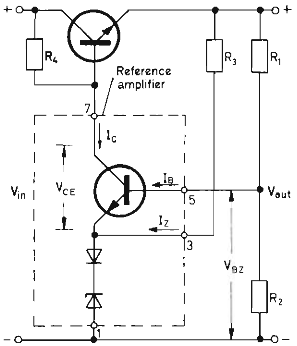


Fig. 20: Schematic of a series voltage regulator circuit in which a reference amplifier is incorporated

2.1.3.2. Thermal resistance

Since the same conditions apply here as in the case of transistors and diodes, this section will be brief. With reference to the German standards DIN 41 785 part 2 and DIN 41 790, we define "thermal resistance" as

R_{thA} Thermal resistance, junction to ambient air

R_{thC} Thermal resistance, junction to case or stud

R_{thS} Thermal resistance, heat sink to ambient air.

These values enable us to calculate whether the diode chip is heated excessly by the overall dissipation P_{tot} encountered during operation. The relevant equations are explained in section 2.1.5.3. Moreover, the numerical value of thermal resistance is required if the thermal differential resistance is to be computed from the temperature coefficient α_{VZ} (section 2.1.3.3.).

2.1.3.3. Differential resistance

The most important property of a Zener diode for the user is the steepness of the operating characteristic (Fig. 13) because the efficiency of the

stabilising circuit of which the Zener diode forms part depends on this factor [22]. The slope is expressed in terms of the differential resistance

$$r_z = \frac{dV_Z}{dI_Z} \tag{4}$$

DIN 41 785, part 2, and DIN 41 790 define, first of all, the inherent differential resistance,

$$r_{zi} = \left(\frac{\partial V_Z}{\partial I_Z}\right)_{T_j} \tag{5}$$

which can be measured for any point of the characteristic at constant junction temperature, by setting the bias by means of the operating current I_Z and by superimposing thereon a low alternating test current at a frequency of 1 kHz, whereupon the drop of alternating voltage across the Zener diode is measured. The quotient of voltage drop to test current is the inherent differential resistance r_{zj}. The 1 kHz test frequency is chosen such that, due to the thermal inertia of the diode chip no junction temperature variations are generated and, on the other hand, the test result cannot be falsified by capacitances or inductances within the Zener diode or in the test set-up. In Fig. 13, the inherent differential resistance is represented by the slope of the solid and the dashed-line characteristics and is virtually inversely proportional to the operating current, see Fig. 21.

Furthermore, the static differential resistance for changing junction temperature and constant ambient temperature is defined in these standards. It is of greater importance for the consideration of the stabilising qualities of a Zener diode than the inherent differential resistance, because variations of the operating current usually take place so slowly

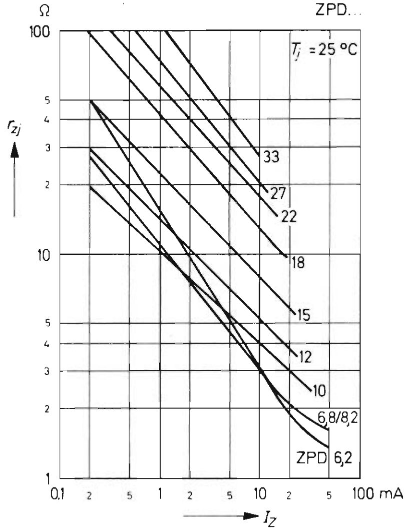


Fig. 21:
Inherent differential resistance versus operating current

that junction temperature varies as well. It is indicated in Fig. 13 by the dotted characteristic. The following equation applies to the static differential resistance [23]

$$r_{zu} = r_{zj} + r_{zth}\,, \tag{6}$$

i. e., the static differential resistance is the sum total of the inherent differential resistance r_{zj} and the thermal differential resistance r_{zth}. The latter represents the temperature coefficient of the operating voltage of the Zener diode and is defined by the equation

$$r_{zth} = \alpha_{VZ} \cdot R_{thA} \cdot V_Z^2 = R_{thA} \cdot V_Z \cdot \frac{\Delta V_Z}{\Delta T_j} \tag{7}$$

wherein

α_{VZ} is the temperature coefficient of the operating voltage,

R_{thA} the thermal resistance junction to ambient air.

Unlike the inherent differential resistance, the thermal differential resistance is virtually independent of operating current.

2.1.3.4. Capacitance

As with any other diode, the reverse-biased PN-junction has capacitance also in the case of Zener diodes [10], [18], such capacitance decreasing as the reverse voltage rises. The capacitance values for $V_R = 1$ V and for $V_R = 2$ V of Zener diodes comprised in the ZPD series can be gathered from the diagram of Fig. 22.

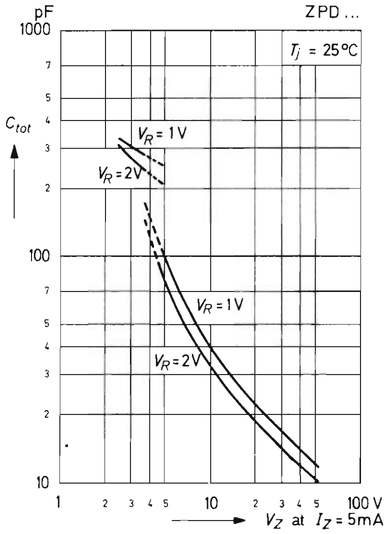


Fig. 22:
Capacitance versus operating voltage

2.1.3.5. Noise

If the noise voltage measured for an avalanche breakdown Zener diode ($V_Z > 7$ V) is plotted as a function of operating current, then a curve with several pronounced maxima is obtained, see Fig. 23. At each noise maximum the current jumps by approximately 100 μA. The mean value of the duty factor can be varied according to the applied voltage. The mean frequency depends, among other things, also on the capacities present in the test circuit.

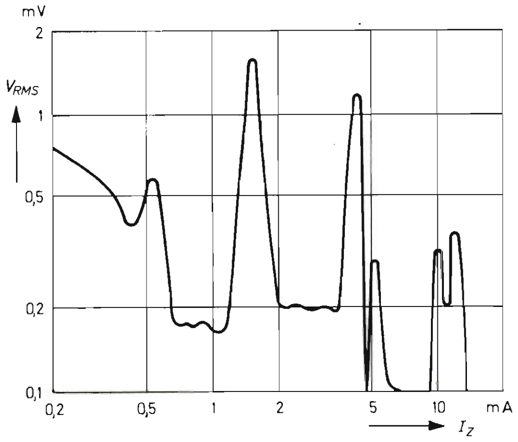


Fig. 23: Noise voltage of a Zener diode versus operating current

When looking for the cause of this fine structure in the breakdown characteristic, typical for all avalanche breakdown Zener diodes, the fact can be exploited that light is emitted upon breakdown. This can be observed if sufficient material above the junction is ground away for the remaining layer to transmit light. A small number of point-like light sources will be noticed. They light up one after the other if the voltage is raised. In those places where avalanche breakdown occurs a physical effect called "microplasm" is produced. Its structure has not been fully discovered. Current density in such a microplasm is extremely high and can amount to 10^4 A/cm^2.

The noise voltage may be reduced by a factor of 10 if a capacitor of about 0.1 μF is connected in parallel with the Zener diode. In contradistinction from glow tube stabilisers, it is permissible for a capacitance to be connected in parallel with Zener diodes without initiating sawtooth oscillations, because the breakdown characteristic does not comprise a negative portion, apart from the above mentioned fine structure. Besides, it is possible to prevent the Zener diode from operating in the region of

the bend in the characteristic where the noise is maximised, if a sufficiently high operating current is chosen. In Zener diodes with operating voltages below 6 V the above described noise is virtually absent, since their characteristic in the region of the bend is steady (Fig. 13).

2.1.4. Measuring the operating voltage of Zener diodes

If a current I_Z traverses a Zener diode, then at constant ambient temperature the operating voltage changes and approaches a final value asymptotically. This voltage changes due to the power dissipated in the junction which in turn causes a rise in junction temperature. Zener diodes with a negative temperature coefficient exhibit a voltage reduction, whereas those with a positive temperature coefficient show a voltage increase on the application of current. The magnitude of the voltage change due to intrinsic heat generation can be derived from the relevant curves quoted in the data book [23].

Because it is not feasible to wait during tests until each device has reached its thermal equilibrium, it is common practice to measure the breakdown voltage of Zener diodes by application of a pulsating current of less than one second duration. Under these conditions the junction temperature remains the same as the ambient temperature. The magnitude of the test current used varies from type to type and is quoted in the data book.

Therefore, designers when devising stabilising circuits and, also, customers carrying out acceptance tests on Zener diodes should allow for the fact that the operating voltage of a device which is at thermal equilibrium will differ from that quoted in the data sheet. In order to asses this voltage, take the operating voltage quoted in the data sheet for $T_j = 25\ ^\circ C$ and, using the method indicated in section 2.1.5.2., calculate the static operating voltage.

2.1.5. Design of stabiliser circuits with Zener diodes

2.1.5.1. Stabilising factor, source resistance

The equations given below apply under the simplifying assumption that the differential resistance of the Zener diode is of constant magnitude. However, since the inherent differential resistance r_{zj} is roughly inversely proportional to the operating current of the Zener diode – as was pointed out in section 2.1.3.3. – and the cooling arrangements for the Zener diode affect the thermal differential resistance, each case needs special consideration if a realistic value for r_z is to be attained. Thus, a value should be chosen for r_{zj} which relates roughly to the centre of a given range of operating currents. Another assumption is that T_{amb} is of constant magnitude. If this is not the case, then the additional fluctuation of the operating voltage may be calculated separately by means of the temperature coefficient.

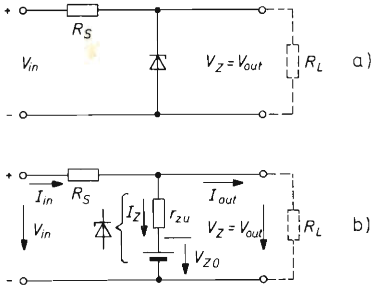


Fig. 24:
Parallel stabiliser circuit with Zener diode
a) Circuit diagram
b) Equivalent circuit

Fig. 24a shows the circuit diagram of a simple parallel stabiliser circuit with a Zener diode, and Fig. 24b the corresponding equivalent circuit in which the Zener diode is replaced by an ideal voltage source in series with a resistor. In this diagram, V_{Z0} is the operating voltage of the Zener diode, extrapolated for zero operating current, and r_{zu} its differential resistance. Other parameters in this circuit diagram are output voltage and output current (V_{out} and I_{out} respectively), input voltage and input current (V_{in} and I_{in} respectively), and operating current I_Z and series resistances R_S. The following equations apply

$$V_{in} = V_{out} + R_S\,(I_{out} + I_Z) \tag{8}$$

and

$$V_{out} = V_{Z0} + I_Z \cdot r_{zu}\,. \tag{9}$$

If equation 9 is written in terms of I_Z and substituted in expression 8, we obtain

$$V_{in} = V_{out} + I_{out} \cdot R_S + (V_{out} - V_{Z0}) \cdot \frac{R_S}{r_{zu}} \tag{10}$$

Assuming I_{out} to be constant, differentiation yields the smoothing factor

$$G = \frac{dV_{in}}{dV_{out}} = 1 + \frac{R_S}{r_{zu}} \approx \frac{R_S}{r_{zu}} \tag{11}$$

For a given Zener diode, this factor increases with R_S. Even an ohmic potential divider may have a considerable smoothing factor, but the percentage fluctuation of input and output voltage is the same. It has therefore proved useful to introduce the ratio of the relative values of input and output voltage changes as the stabilising factor S. From equation 11 we obtain:

$$S = \frac{\dfrac{dV_{in}}{V_{in}}}{\dfrac{dV_{out}}{V_{out}}} = \left(1 + \frac{R_S}{r_{zu}}\right)\frac{V_{out}}{V_{in}} \approx \frac{R_S}{r_{zu}} \cdot \frac{V_{out}}{V_{in}} \tag{12}$$

The stabilisation factor, unlike the smoothing factor, does not increase linearly with V_{in} and R_S, but approaches a finite limit S_{max} which is calculated as follows: R_S is eliminated from equation 12 by using equation 8:

$$R_S = \frac{V_{in} - V_{out}}{I_{out} + I_Z} = \frac{V_{in} - V_{out}}{I_{in}} \tag{13}$$

$$S = \frac{V_{out}}{V_{in}} + \frac{V_{in} - V_{out}}{I_{in} \cdot r_{zu}} \cdot \frac{V_{out}}{V_{in}} = \frac{V_{out}}{V_{in}} + \frac{V_{out}}{I_{in} \cdot r_{zu}} \cdot \left(1 - \frac{V_{out}}{V_{in}}\right) \tag{14}$$

By substituting $V_{in} = \infty$, we obtain

$$S_{max} = \frac{V_{out}}{I_{in} \cdot r_{zu}} \tag{15}$$

It can be seen that for a given Zener diode and a given load, stabilisation improves as the input voltage is increased; it should be noted, however, that the power dissipated in the series resistor rises at a rate higher than that at which the stabilisation factor is increased. As a sensible compromise between the requirements of good stabilisation and acceptable power dissipation, it is suggested that the input voltage be made about two to four times the value of the output voltage. If the internal resistance of the voltage source V_{in} is neglected, we obtain the output resistance of the circuit illustrated in Fig. 24 from the parallel connection of r_{zu} and R_S:

$$r_{out} = \frac{r_{zu} \cdot R_S}{r_{zu} + R_S} \tag{16}$$

and, by approximation (since, as a rule, $R_S \gg r_{zu}$),

$$r_{out} \approx r_{zu} \, . \tag{17}$$

2.1.5.2. Operating voltage, operating characteristic

As explained in section 2.1.4. the operating voltage quoted on the data sheet or in the data book [23] for every type of the Zener diode series is tested for a test current I_{Ztest} with pulses of a duration short enough for the junction temperature of the diode to remain the same as ambient temperature. By the same method the operating characteristics of which Fig. 25 shows an example are likewise determined point by point. Usually the temperature dependence of the operating voltage, demonstrated in Fig. 13, is not expressed in such a family of curves, nor is Fig. 25 suitable for giving indications of the spread of operating voltages and differential resistance. This information is contained in the table of data, shown below by way of example, and in some further diagrams.

With reference to the Zener diode ZPD 10, we shall now discuss a simple parallel stabilising circuit according to Fig. 24, assuming a nominal input voltage of 24 V which may fluctuate by ± 4 V, and a load resistance R_L in the region of 500 Ω . . . 2 kΩ. The admissible power dissipation of the ZPD 10 is 500 mW.

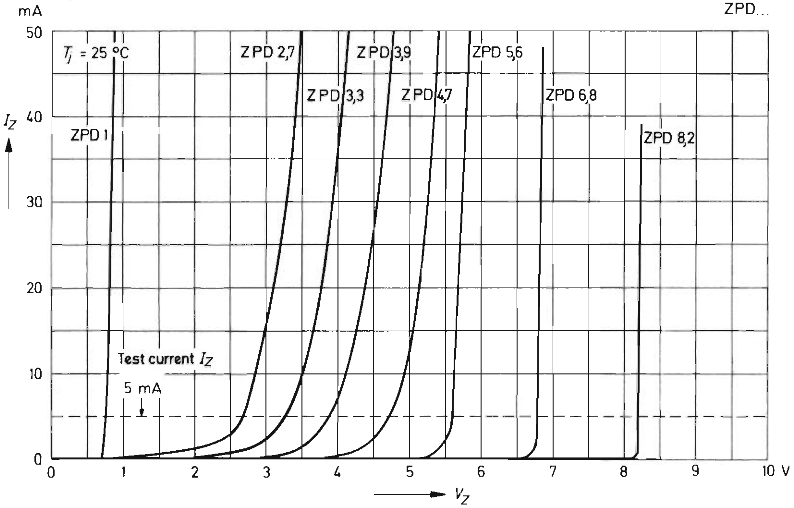


Fig. 25: Operating characteristics of Zener diodes (tested with pulses)

The series resistance R_S is subject to two criteria: when the input voltage V_{in} and the load resistance R_L are at their lowest an adequate operating current should still flow through the Zener diode; on the other hand, when the input voltage and the load resistance are at their highest, the admissible power dissipation of the Zener diode should not be exceeded. For the former case we assume a Zener diode operating at the upper limit of the V_Z spread, $V_{Z\,max} = 10.6$ V, considering a minimum operating current $I_{Z\,min} = 3$ mA to be adequate. We obtain the following expressions:

$$I_{in} = \frac{V_{in\,min} - V_{Z\,max}}{R_S} > I_{Z\,min} + I_{out\,max} \tag{18}$$

or, resolved in terms of R_S

$$R_S < \frac{V_{in\,min} - V_{Z\,max}}{I_{Z\,min} + I_{out\,max}} \tag{19}$$

and, substituting numerical values

$$R_S < \frac{20\text{ V} - 10.6\text{ V}}{3\text{ mA} + (10.6\text{ V} : 500\ \Omega)} = \frac{9.4\text{ V}}{3\text{ mA} + 21\text{ mA}} = 390\ \Omega\,.$$

The least favourable situation as regards dissipation obtains in the case of a Zener diode at the lower limit of the spread when $V_{Z\,min} = 9.4$ V, for a maximum input voltage $V_{in\,max} = 28$ V and a maximum load resistance $R_L = 2$ kΩ. The relationship should be

$$V_{Z\,min} \cdot I_{Z\,max} < P_{tot}\,, \tag{20}$$

Table 2:
Main data of ZPD 1 . . . ZPD 33 Zener diodes

Type	Operating voltage [1]) @ I_Z = 5 mA	Dynamic resistance @ I_Z = 5 mA f = 1 kHz	@ I_Z = 1 mA f = 1 kHz	Temp. coeff. of Oper. volt. @ I_Z = 5 mA	Reverse voltage @ I_R = 100 nA	Admissible Zener current [2]) @ T_{amb} = 45 °C	@ T_{amb} = 25 °C
	V_Z V	r_{zj} Ω	r_{zj} Ω	α_{VZ} 10^{-4}/°C	V_R V	I_Z mA	I_Z mA
ZPD 1 [3])	0.7...0.8	6.5 (<8)	<50	–26...–23	–	280	340
ZPD 2,7	2.5...2.9	75 (<83)	<500	–9...–4	–	135	160
ZPD 3	2.8...3.2	80 (<95)	<500	–9...–3	–	117	140
ZPD 3,3	3.1...3.5	80 (<95)	<500	–8...–3	–	109	130
ZPD 3,6	3.4...3.8	80 (<95)	<500	–8...–3	–	101	120
ZPD 3,9	3.7...4.1	80 (<95)	<500	–7...–3	–	92	110
ZPD 4,3	4.0...4.6	80 (<95)	<500	–6...–1	–	85	100
ZPD 4,7	4.4...5.0	70 (<78)	<500	–5...+2	–	76	90
ZPD 5,1	4.8...5.4	30 (<60)	<480	–3...+4	>0.8	67	80
ZPD 5,6	5.2...6.0	10 (<40)	<400	–2...+6	>1	59	70
ZPD 6,2	5.8...6.6	4.8 (<10)	<200	–1...+7	>2	54	64
ZPD 6,8	6.4...7.2	4.5 (<8)	<150	+2...+7	>3	49	58
ZPD 7,5	7.0...7.9	4 (<7)	<50	+3...+7	>5	44	53
ZPD 8,2	7.7...8.7	4.5 (<7)	<50	+4...+7	>6	40	47
ZPD 9,1	8.5...9.6	4.8 (<10)	<50	+5...+8	>7	36	43
ZPD 10	9.4...10.6	5.2 (<15)	<70	+5...+8	>7.5	33	40
ZPD 11	10.4...11.6	6 (<20)	<70	+5...+9	>8.5	30	36
ZPD 12	11.4...12.7	7 (<20)	<90	+6...+9	>9	28	32
ZPD 13	12.4...14.1	9 (<25)	<110	+7...+9	>10	25	29
ZPD 15	13.8...15.6	11 (<30)	<110	+7...+9	>11	23	27
ZPD 16	15.3...17.1	13 (<40)	<170	+8...+9.5	>12	20	24
ZPD 18	16.8...19.1	18 (<50)	<170	+8...+9.5	>14	18	21
ZPD 20	18.8...21.2	20 (<50)	<220	+8...+10	>15	17	20
ZPD 22	20.8...23.3	25 (<55)	<220	+8...+10	>17	16	18
ZPD 24	22.8...25.6	28 (<80)	<220	+8...+10	>18	13	16
ZPD 27	25.1...28.9	30 (<80)	<250	+8...+10	>20	12	14
ZPD 30	28...32	35 (<80)	<250	+8...+10	>22.5	10	13
ZPD 33	31...35	40 (<80)	<250	+8...+10	>25	9	12

1) tested with pulses

2) Valid provided that the leads are kept at ambient temperature at a distance of 8 mm from case.

3) The ZPD 1 is a silicon diode with operation in forward direction. Hence, the index of all parameters should be "F" instead of "Z". Connect the cathode lead to the negative pole.

and with equation

$$I_{Z\,max} = \frac{V_{in\,max} - V_{Z\,min}}{R_S} - I_{out\,min} \tag{21}$$

we obtain

$$V_{Z\,min} \cdot \left(\frac{V_{in\,max} - V_{Z\,min}}{R_S} - I_{out\,min}\right) < P_{tot}\,, \tag{22}$$

hence

$$R_S > \frac{V_{Z\,min} \cdot V_{in\,max} - (V_{Z\,min})^2}{P_{tot} + V_{Z\,min} \cdot I_{out\,min}} \tag{23}$$

and, by substituting numerical values

$$R_S > \frac{9.4\text{ V} \cdot 28\text{ V} - (9.4\text{ V})^2}{500\text{ mW} + 9.4\text{ V} \cdot (9.4\text{ V}/2\text{ k}\Omega)}$$

$$= \frac{263\text{ V}^2 - 88\text{ V}^2}{500\text{ mW} + 44\text{ mW}} = \frac{175\text{ V}^2}{0.544\text{ W}} = 322\ \Omega\,.$$

The series resistance R_S to be selected will have a value of 360 Ω, with a tolerance of ± 10 %. It should be rated for the following power dissipation:

$$P = \frac{(V_{in\,max} - V_{Z\,min})^2}{R_S} \tag{24}$$

or, numerically

$$P = \frac{(28\text{ V} - 9.4\text{ V})^2}{360\ \Omega} = 0.96\text{ W} \approx 1\text{ W}\,.$$

Having determined the value for R_S, it is possible to estimate the region in which the stabilised voltage can be found. For V_{in} = 20 V, V_Z = 9.4 V at I_Z = 5 mA and R_L = 500 Ω, the Zener diode is traversed by an operating current of approximately

$$I_Z = \frac{V_{in} - V_Z}{R_S} - \frac{V_Z}{R_L}\,, \tag{25}$$

i. e.

$$I_Z = \frac{20\text{ V} - 9.4\text{ V}}{360\ \Omega} - \frac{9.4\text{ V}}{500\ \Omega} = 29.4\text{ mA} - 18.8\text{ mA} \approx 11\text{ mA}.$$

This value allows the output voltage V_{out} to be determined if the operating voltage V_{Z0}, which has been extrapolated for I_Z = 0, is assumed to be 9.2 V and the inherent differential resistance for I_Z = 5 mA is taken from Fig. 21 to be r_{zj} = 5 Ω and the thermal differential resistance is taken from Fig. 26 to be r_{zth} = 20 Ω (equation 9):

$$V_{out} = V_{Z0} + I_Z \cdot r_{zu} = 9.2\text{ V} + 11\text{ mA} \cdot (5\ \Omega + 20\ \Omega)$$

$$= 9.2\text{ V} + 0.275\text{ V} = 9.475\text{ V} \approx 9.5\text{ V}.$$

If the ZPD 10 device under consideration has an operating voltage of V_Z = 10.6 V and if we assume V_{Z0} to be 10.4 V and if, this time, we calculate

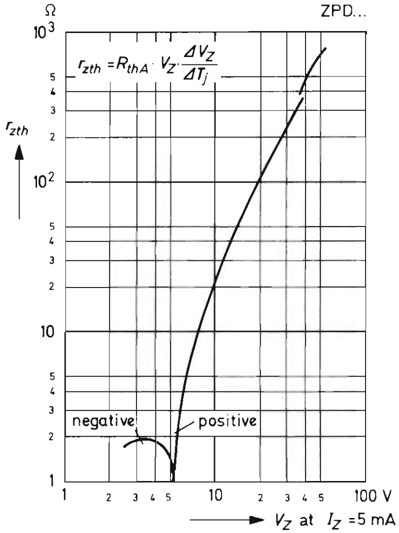


Fig. 26: Thermal differential resistance versus operating voltage

not with the previously substituted typical value for r_{zj}, but with a value three times higher (at the upper limit of the spread, see table 2), and if we substitute for R_L and V_{in} the upper values of 2 kΩ and 28 V respectively, then the Zener diode operates with a current

$$I_Z = \frac{28\ \text{V} - 10.6\ \text{V}}{360\ \Omega} - \frac{10.6\ \text{V}}{2\ \text{k}\Omega} = 48.3\ \text{mA} - 5.3\ \text{mA} = 43\ \text{mA}$$

and, if we estimate r_{zj} from Fig. 21 to be approximately 6 Ω, the output voltage V_{out} of the stabiliser circuit (Fig. 24) amounts to:

$$V_{out} = V_{Z0} + I_Z \cdot r_{zu} = 10.4\ \text{V} + 43\ \text{mA} \cdot (6\ \Omega + 20\ \Omega)$$

$$= 10.4\ \text{V} + 1.12\ \text{V} = 11.52\ \text{V} \approx 11.5\ \text{V}.$$

In order to perform an exact calculation, we should now actually recalculate the equations for I_Z, using the values for $V_{out} = V_Z$ of 9.5 V and 11.5 V respectively, in order to determine the corrected values for V_{out}. However, the improved accuracy of the calculation would be illusory, for the values for r_z used in such a calculation could only be ascertained approximately.

However, the dissipation of the heat generated in the Zener diode, and the effect of cooling upon thermal resistance must still be considered. This will be discussed later on.

Finally, we are concerned with ascertaining the stabilising factor for the circuit under consideration. For a mean input voltage $V_{in} = 24$ V and a

mean load resistance $R_L = 1\ \text{k}\Omega$, the Zener diode (a typical ZPD 10) conducts an operating current

$$I_Z = \frac{V_{in} - V_Z}{R_S} - \frac{V_Z}{R_L} = \frac{24\ \text{V} - 10\ \text{V}}{360\ \Omega} - \frac{10\ \text{V}}{1\ \text{k}\Omega}$$

$$= 39\ \text{mA} - 10\ \text{mA} = 29\ \text{mA}.$$

With this current, the inherent dynamic resistance r_{zi} has a value of typically 2.5 Ω (Fig. 23). Equation 12 reveals the stabilising factor

$$S = \left(1 + \frac{R_S}{r_{zu}}\right) \cdot \frac{V_{out}}{V_{in}} = \left(1 + \frac{360\ \Omega}{2.5\ \Omega + 20\ \Omega}\right) \cdot \frac{10\ \text{V}}{24\ \text{V}}$$

$$= (1 + 16) \cdot 0.42 = 7.1\ .$$

Thus the stabilising circuit illustrated in Fig. 24 is capable of reducing the input voltage fluctuations assumed to be 24 V ± 4 V (corresponding to ± 17 %) to one seventh of 17 %, i. e. ± 2.5 %. This applies to a situation with constant load resistance and constant ambient temperature. If the latter fluctuates, too – and this will usually be the case –, then the influence of ambient temperature on the output voltage V_{out} can be determined with the aid of Figs. 14 or 17 or with the aid of the temperature coefficient quoted in table 2.

From the calculations carried out so far it is apparent that the thermal differential resistance r_{zth} is about four to eight times higher than the inherent differential resistance r_{zi} (depending upon current) and, thus, has a considerable effect upon the stabilising factor *S*. It may therefore be advisable to employ, instead of the 500 mW Zener diode ZPD 10, a 1 watt Zener diode ZPY 10. The thermal differential resistance r_{zth} of the latter may be calculated for $R_{thA} = 130\ °\text{C/W}$ and for $\alpha_{VZ} = 7 \cdot 10^{-4}/°\text{C}$ according to equation 7:

$$r_{zth} = \alpha_{VZ} \cdot R_{thA} \cdot V_Z^2 = 7 \cdot 10^{-4}\ °\text{C}^{-1} \cdot 130\ °\text{C/W} \cdot 100\ \text{V}^2 = 9.1\ \Omega\ ,$$

this being half the value for ZPD 10 (20 Ω). An even lower value, i. e. 0.35 . . . 5.7 Ω (depending upon the size of the cooling area) can be obtained with the power Zener diode ZX 10.

If the operating current of the Zener diode fluctuates rapidly, e. g. as a result of a 100 Hz hum superimposed upon the input voltage V_{in}, then only the inherent differential resistance r_{zi} needs to be substituted in order to ascertain the stabilising factor, whereby a stabilising factor roughly eight times higher, i. e. about 56, results for the circuit under discussion, in relation to the hum voltage.

We conclude this section by stating that there is no simple formula by means of which all parameters of a stabilising circuit can be calculated quickly and accurately. Rather, several assumptions have to be made to begin with, and a Zener diode type must be selected. This is followed by calculations and by correcting the initial assumptions if necessary, by recalculating and so forth, until the circuit has been dimensioned and characterised satisfactorily. Never strive for absolute accuracy, because all the parameters of semiconductor devices are subject to spread and,

moreover are frequently dependent on current and temperature. The result of a calculation is affected at least by those errors which attach to numerical values substituted in the equations. Besides, it would rarely be advisable to aim at dimensioning a circuit according to Fig. 22 for an output voltage of exactly 10 V, because commercial Zener diodes operate with a voltage tolerance of ± 5 %. In order to obtain 10 V exactly it would be better to employ a 12 V Zener diode ZPD 12 and to set the stipulated 10 V with the aid of a potentiometer connected to the stabiliser circuit.

For special applications which do not require devices in large numbers it would also be possible to buy Zener diodes with closely tolerated operating voltage e. g. ± 2 % or ± 1 %. However, since such selection may necessitate considerable surcharges, it will be advisable to check in every case whether the setting of the voltage with the aid of a potentiometer or the employment of a closely tolerated Zener diode is the more economical proposition.

2.1.5.3. Power dissipation, thermal resistance, cooling measures

If a Zener diode is biased with direct current, a power dissipation P_{tot} occurs in the silicon chip which is the product of operating voltage and operating current:

$$P_{tot} = V_Z \cdot I_Z \qquad (26)$$

By taking suitable measures – chosing a moderate operating current, an amply rated Zener diode and by taking care to ensure satisfactory dissipation of heat, by an additional cooling fin if necessary – the designer must make sure that the admissible junction temperature is not exceeded under any of the possible operating conditions [24], [25], [26]. Quite apart from this, it may be advantageous – as pointed out in the preceding section – to use a Zener diode of thermally ample proportions because this increases the stabilising factor.

After ascertaining the worst possible power dissipation encountered during operation by means of equation 26, it is possible to calculate the resulting junction temperature by means of the thermal resistance R_{thA} or R_{thC} quoted in the data sheet or data book [23]. The junction temperature becomes

$$T_j = T_{amb} + P_{tot} \cdot R_{thA} \qquad (27)$$

This expression applies, for example, to glass diodes of the ZPD . . ., ZPY . . . and ZPU . . . series as well as the power Zener diodes of type ZX . . . without cooling fin. When substituting the value of the thermal resistance the fact that especially in the case of glass diodes a considerable proportion of the heat generated in the silicon crystal is dissipated via the connection leads should be taken into account. Therefore, the thermal resistance effective in the case under consideration depends on the distance between the case and the point at which the leads are soldered to an element acting as heat sink, e. g. the circuit board (see also the curves shown in Fig. 27). The temperature at the soldering point encountered during operation will also depend on other heat generators

connected to the circuit board and should be inserted in equation 27 as the ambient temperature T_{amb}.

In the case of power Zener diodes with threaded case (Fig. 12), the total effective thermal resistance is made up of the inner thermal resistance R_{thC} which becomes effective between the chip and the base (or screw) of the diode case, and the thermal resistance R_{thS} of the cooling fin or heat sink, which is effective between the point at which the diode is mounted and the air surrounding the heat sink. In this way, equation 27 is expanded to read:

$$T_j = T_{amb} + P_{tot} \cdot (R_{thC} + R_{thS}) \qquad (28)$$

Under certain conditions, and especially when the Zener diode is mounted in an isolated position, the total thermal resistance, comprised in parentheses, is increased further by the thermal resistance diode to cooling fin, and in the case of insulated mounting with a mica washer of approximately 50 μm thickness, by approximately 1 °C/W. If the required cooling fin area is to be calculated when the maximum ambient temperature and the power dissipation are given, then we obtain from equation 28

$$R_{thS} = \frac{T_j - T_{amb}}{P_{tot}} - R_{thC} \qquad (29)$$

and the required minimum size of the cooling fin can then be ascertained from Fig. 28 [23].

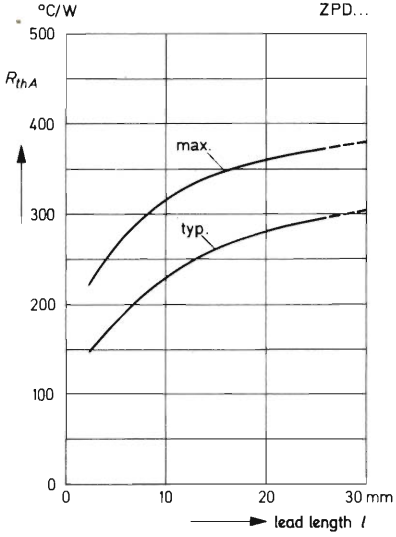


Fig. 27:
Thermal resistance versus lead length

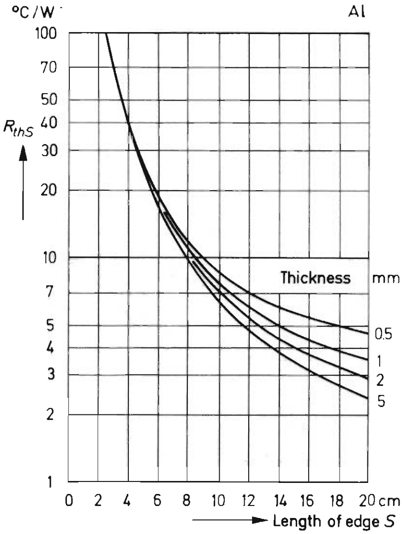


Fig. 28:
Diagram for ascertaining the thermal resistance of Al-cooling fins as a function of the length of edge, S

The characteristics are valid for bare, vertically positioned, approximately square cooling fins with a centrally mounted semiconductor device, in still air and without additional irradiation of heat. For horizontal arrangements, the calculated area of the cooling fin should be multiplied by a factor of 1.3. If the cooling fin is blackened, its area may be reduced in relation to the calculated area by a factor of 0.7.

The above information applies to constant power dissipation and a constant operating current I_Z (direct current). There are applications however, when the operating current is not continuous, the simplest case being that of a periodic squarewave pulse, for example. Depending upon the frequency f_p and the duty cycle ν of such pulses, there are many cases in which the admissible current rating is much higher than that stated for the admissible static power dissipation. The operating current admissible during pulse increases with a decreasing duty cycle. The thermal capacity of the diode system and the continuous dissipation of heat prevent the pulsed power dissipation causing unduly high junction temperatures. The thermal conditions under pulsed loading can be calculated by means of the family of curves shown in Fig. 29 [23].

From Fig. 29, the "pulse thermal resistance" r_{thA} can be gathered as a function of the pulse duration t_p and the duty cycle. For the squarewave power dissipation P_I we obtain a junction temperature

$$T_j = T_{amb} + P_I \cdot r_{thA} \,. \qquad (30)$$

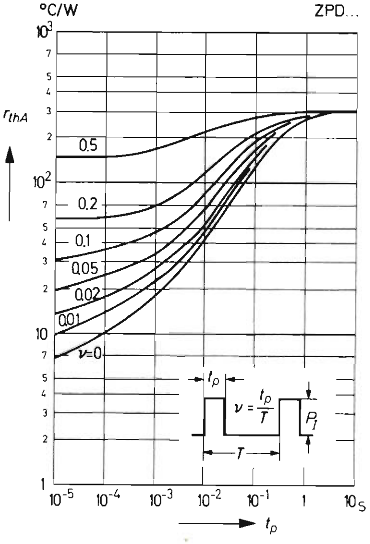


Fig. 29:
Pulse thermal resistance versus pulse duration

If the pulse mode dissipation P_I is superimposed on a continuous power dissipation P_D, then the expression 30 will have to be expanded to read

$$T_j = T_{amb} + P_D \cdot R_{thA} + P_I \cdot r_{thA} . \tag{31}$$

The expression

$$T_j = T_{amb} + P_{tot} \cdot R_{thS} + P_I \cdot r_{thC} \tag{32}$$

applies to a power Zener diode ZX... screw-connected to a cooling sheet. In this equation P_{tot} is the mean of the power dissipation P_I. If to this must be added the continuous power dissipation P_D, we obtain

$$T_j = T_{amb} + P_{tot} \cdot R_{thS} + P_D \cdot R_{thC} + P_I \cdot r_{thC} \tag{33}$$

wherein P_{tot} is the mean value of the total power dissipation.

The pulse thermal resistance r_{thA} indicated in Fig. 29, like the static thermal resistance R_{thA}, is dependent on the lenght of the connection leads between the diode case and the soldering point kept at ambient temperature. Therefore, the value for r_{thA}, derived from Fig. 29, may have to be corrected according to the curve in Fig. 27.

2.2. Stabiliser diodes

2.2.1. The structure of stabiliser diodes

For several years, INTERMETALL has supplied two stabiliser diodes with operating voltages of 1.5 and 2 V at a current of 5 mA, carrying the type designations ZE 1,5 and ZE 2 (in a plastic case, see Fig. 30). These were in fact two (three) series-connected diode chips, operating in the forward direction, which were comprised in a drop-shaped plastic case. As mentioned before in section 1.2.7., stabiliser diodes of this type were employed for stabilising voltages below 3 V, since for this range useful Zener diodes were not available.

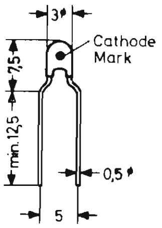


Fig. 30:
Stabiliser diode ZE 1,5 or ZE 2
in a plastic case (1966 . . .1974)

In the meantime, INTERMETALL has developed and marketed an improved series of "stabiliser diodes" by launching the type ZTE 1,5 . . . ZTE 5,1 series of "stabiliser diodes". These devices, which are supplied in a glass case DO-35 (Fig. 11), are in fact integrated analog circuits. Fig. 31 shows the internal circuit of types ZTE 1,5 and ZTE 2, and Fig. 32 shows that of types ZTE 2,4 . . . ZTE 5,1. From Fig. 31 can be gathered that the internal circuit of types ZTE 1,5 and ZTE 2 is extremely simple, i. e. a Darlington circuit, the operating voltage being equal to the sum of the two or three base emitter voltages, i. e. approximately 1.4 V and 2.1 V respectively. In order to achieve higher operating voltage, it would be possible to combine a suitably increased number of transistors to form a Darlington circuit; however, another circuit – Fig. 32 – has proved more economic and technically more advantageous.

The circuit of Fig. 32 consists of an adjustable reference voltage source comprising T3, R_2 and R_3 [27], [28], which is followed by the Darlington circuit made up of transistors T1 and T2. This circuit is applicable to all the types ZTE 2,4 . . . ZTE 5,1. Besides, these types are also diffused together. The desired operating voltage is then set in common for all elements of a wafer by adjusting resistor R_3 by means of the aluminium wiring. The advantage of an adjustable reference voltage source is brought about by the fact that any desired operating voltage can be set by adjusting R_3, whereas in a circuit according to Fig. 31 only integral multiples of 0.7 V are feasible.

A common feature of both the circuits of Fig. 31 and Fig. 32 is the additional diode connected in antiparallel configuration to the forward bias.

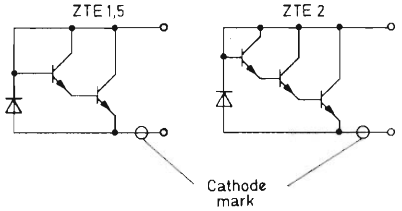


Fig. 31:
Internal circuit of
ZTE 1,5 and ZTE 2

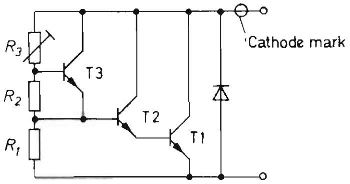


Fig. 32:
Internal circuit of
ZTE 2,4 . . . ZTE 5,1

This diode produces a well defined behaviour in the reverse direction, i. e. the behaviour of a forward biased diode. This is why ZTE stabilising diodes resemble normal Zener diodes also in the reverse direction.

2.2.2. Data of stabiliser diodes ZTE 1,5 . . . ZTE 5,1

The table given below contains data of the ZTE type series.

Table 3:
Main data of the stabiliser diodes ZTE 1,5 . . . ZTE 5,1

Type	Operating voltage @ $I_Z = 5$ mA	Dynamic resistance @ $I_Z = 5$ mA	permissible operating current @ $T_{amb} = 25$ °C [1]
	V_Z V	r_{zi} Ω	I_Z max. mA
ZTE 1,5	1.35...1.55	13 (< 20)	120
ZTE 2	2.0 ...2.3	18 (< 30)	120
ZTE 2,4	2.2 ...2.56	14 (< 20)	120
ZTE 2,7	2.5 ...2.9	15 (< 20)	105
ZTE 3	2.8 ...3.2	15 (< 20)	95
ZTE 3,3	3.1 ...3.5	16 (< 20)	90
ZTE 3,6	3.4 ...3.8	16 (< 25)	80
ZTE 3,9	3.7 ...4.1	17 (< 25)	75
ZTE 4,3	4.0 ...4.6	17 (< 25)	65
ZTE 4,7	4.4 ...5.0	18 (< 25)	60
ZTE 5,1	4.8 ...5.4	18 (< 25)	55

[1]) Valid provided that the leads are kept at ambient temperature at a distance of 8 mm from case.

A comparison with the data of the ZPD series, contained in table 2 (page 32) shows that the inherent differential resistance of the ZTE diodes is considerably lower. From the operating characteristics of Fig. 33 this becomes also apparent if a comparison with Fig. 25 is made. A striking feature of table 3 is the comparatively large negative temperature coefficient of the operating voltage. However, that which seems a disadvantage to begin with proves an advantage, as will now be explained with reference to Fig. 34 (see also Fig. 26 for comparison). In Fig. 34, the two components of the static differential resistance r_{zu}, the inherent differential resistance r_{zi} and the thermal differential resistance r_{zth} are shown. If r_{zi} and r_{zth} are added according to equation 6, in order to obtain r_{zu}, then we obtain, as the sum total of a positive (r_{zi}) and a negative (r_{zth}) resistance, the difference of the two numerical values and thus an especially low value for r_{zu}, which usually determines the stabilising factor. Below, a numerical example:

At $I_Z = 5$ mA, a Zener diode ZPD 4,7 has a static differential resistance of

$$r_{zu} = r_{zi} + r_{zth} = 70\ \Omega - 1.7\ \Omega = 68.3\ \Omega \approx 70\ \Omega\ ,$$

i. e., the inherent differential resistance (gathered from table 2) of 70 Ω is so large that the thermal differential resistance (obtained from Fig. 26), which amounts to about -1.7 Ω can be neglected. For the stabiliser diode ZTE 4,7, on the other hand, we obtain at $I_Z = 5$ mA

$$r_{zu} = r_{zi} + r_{zth} = 18\ \Omega - 13\ \Omega = 5\ \Omega\ .$$

This considerable advantage of the ZTE 4,7 applies only at constant ambient temperature. If the latter fluctuates, the temperature-dependent variations of the operating voltage are greater in the case of the ZTE 4,7

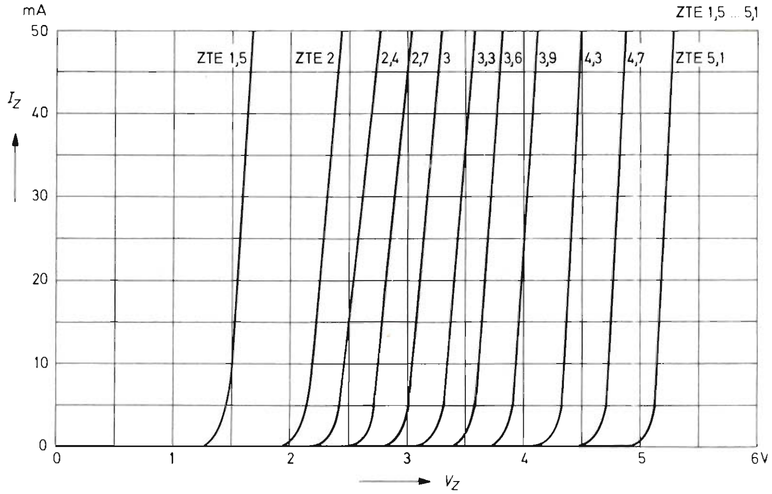


Fig. 33: Operating characteristics of the ZTE diodes (tested with pulses)

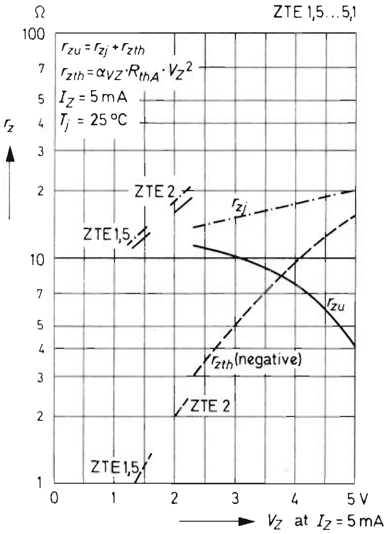


Fig. 34:
Differential resistance versus operating voltage

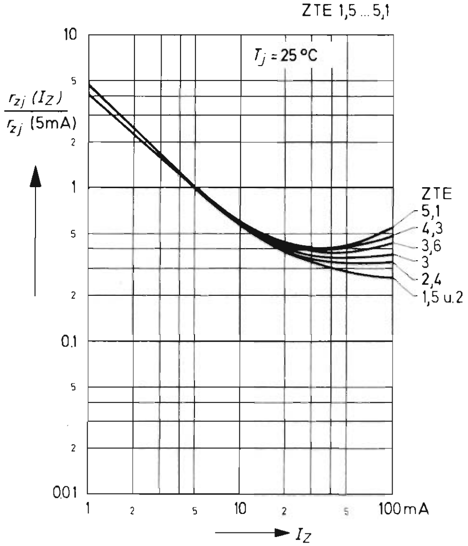


Fig. 35:
Inherent differential resistance versus operating current (relative values)

than in the case of the ZPD 4,7, and one will have to decide whether in that case the employment of the ZPD 4,7 is perhaps more advantageous, notwithstanding the high inherent differential resistance. In this context, Fig. 35 shows also the dependence of inherent differential resistance on the operating current. Otherwise, the explanations given in section 2.1.5. apply also to the ZTE diodes.

2.3. Temperature-compensated Zener diodes

Attempts were made at an early stage to replace by Zener diodes the standard cells [29] then used as voltage reference sources. These cells were heavy and expensive and of limited usefulness. In 1960, INTERMETALL presented the silicon reference elements BZY 22 . . . 25 [20] which, at best, operated with a temperature coeffizient of the operating voltage $\alpha_{VZ} < 10^{-5}/°C$ for a current of 5 mA. The operating voltage amounted to 8.4 V. However, the manufacture of these reference elements was very labour-intensive, so that in 1967 the time was considered ripe for integrated components, i. e. temperature-compensated Zener diodes. At first, INTERMETALL put the ZTK 33 on the market [30] which was developed specifically for stabilising the tuning voltage of diode operated TV tuners [18]. Later developments resulted in a whole series of temperature-compensated Zener diodes, comprising the types ZTK 6,8 . . . ZTK 33.

2.3.1. Methods of temperature compensation

It will be gathered from Fig. 17 that temperature compensation of the operating voltage of Zener diodes, achieved by the series connection of forward-biased silicon diodes makes sense only if the temperature coefficient of the Zener diode is independent of temperature, i. e. if the characteristic of ΔV_Z versus T_j in Fig. 17 is a straight line. This is the case for $V_Z > 6$ V [20]. There are essentially three methods of temperature compensation which are shown diagrammatically in Fig. 36 [11].

Fig. 36a shows a series circuit of discrete semiconductor components such as were used in these reference elements. The operating characteristic of this circuit arrangement differs only little from that of a simple Zener diode, but suffers from the disadvantage of a relatively high inherent differential resistance, since the resistance values of the several diodes are added to each other. The thermal coupling between the diodes, and thus the dynamic component of the temperature compensation is not particularly good.

Improved properties can be achieved with discrete semiconductor components in circuit Fig. 36b which contains a shunt transistor that may also be designed as a Darlington circuit. Due to the current gain of the transistor, a small inherent differential resistance is achieved in this arrangement. Here, too, bad thermal coupling is a disadvantage.

The best method is the integration of the shunt transistor circuit, as outlined in Fig. 36c. Here, the shunt transistor ensures that the inherent differential resistance remains very low, and since the complete circuit can be accomodated on an extremely small 0.5 x 0.5 mm^2 silicon chip, good thermal coupling between all components is ensured. In Fig. 36c, the Zener diode is realised by the emitter diode of a transistor operating under reverse breakdown conditions, with an operating voltage of approximately 6 V.

The operating characteristic of the circuits illustrated in Fig. 36b and c differs in one significant respect from that of a conventional Zener diode

(Fig. 36a). There is no sudden rise in current when the voltage applied to the circuit terminals in Fig. 36c reaches the breakdown level of the emitter diode of the transistor acting as a Zener diode. Instead, it rises according to a characteristic which takes resistor *R* into account. It is only when base current begins to flow in the shunt transistor as a result of the voltage increasing further that the characteristic rises steeply and merges into the actual operating region.

2.3.2. The construction of ZTK diodes

All temperature-compensated monolithic integrated Zener diodes of the series ZTK 6,8 . . . ZTK 33, developed by INTERMETALL, are based on the simple shunt transistor circuit of Fig. 36c. In devices designed for higher output voltages than 7 V, several transistors operating as Zener diodes are connected in series each of which, in theory, has to be temperature-compensated by means of one or two forward-biased diodes connected in series thereto. Instead of such diodes, the ZTK diodes comprise a so-called "adjustment stage" of a kind already described

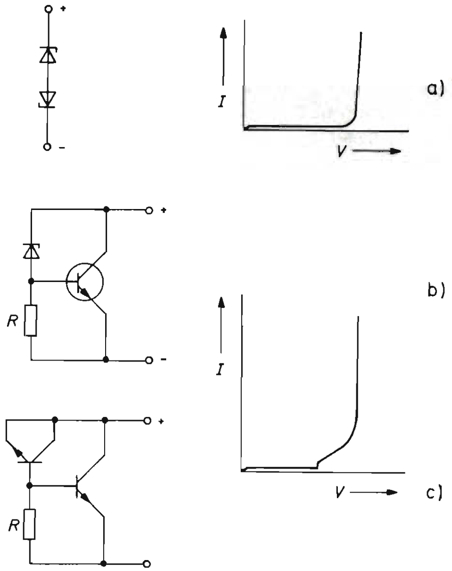


Fig. 36: Methods of temperature compensation
a) Series circuit of several discrete diodes
b) Circuit with Zener diode and shunt transistor, made up of discrete semiconductors
c) Monolithic integrated circuit with Zener diode and shunt transistor

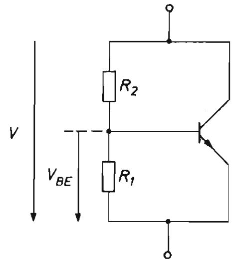


Fig. 37: Adjustment stage

briefly with reference to Fig. 32. Fig. 37 illustrates once more the circuit diagram of such an adjustment stage [27], [28] to which the following expression applies, provided the transistor operates with a sufficiently high current gain.

$$V = V_{BE} \cdot \frac{R_1 + R_2}{R_1} = V_{BE} \cdot \left(1 + \frac{R_2}{R_1}\right) \qquad (34)$$

If the base emitter voltage V_{BE} changes as a result of temperature variations, the voltage V also changes, according to the following expression:

$$\frac{dV}{dT} = \frac{R_1 + R_2}{R_1} \cdot \frac{dV_{BE}}{dT}; \qquad (35)$$

for in a monolithic integrated circuit the ratio of resistors is virtually independent of temperature. With the aid of this adjustment stage, a voltage V is produced which may be any desired – not necessarily integral – multiple of a base emitter voltage V_{BE}, its temperature coefficient being equal to the temperature coefficient of the base emitter voltage of a transistor. This ensures a temperature compensation more accurate than that achievable by means of several diodes in a simple series circuit in which fractions of V_{BE} cannot be realised.

Referring to type ZTK 22, the internal circuit of a ZTK diode (Fig. 38) will now be discussed. Two transistors operating as Zener diodes, each with

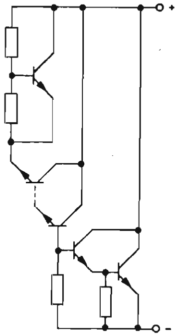


Fig. 38: Internal circuit of the ZTK 22

(Fig. 36a). There is no sudden rise in current when the voltage applied to the circuit terminals in Fig. 36c reaches the breakdown level of the emitter diode of the transistor acting as a Zener diode. Instead, it rises according to a characteristic which takes resistor R into account. It is only when base current begins to flow in the shunt transistor as a result of the voltage increasing further that the characteristic rises steeply and merges into the actual operating region.

2.3.2. The construction of ZTK diodes

All temperature-compensated monolithic integrated Zener diodes of the series ZTK 6,8 . . . ZTK 33, developed by INTERMETALL, are based on the simple shunt transistor circuit of Fig. 36c. In devices designed for higher output voltages than 7 V, several transistors operating as Zener diodes are connected in series each of which, in theory, has to be temperature-compensated by means of one or two forward-biased diodes connected in series thereto. Instead of such diodes, the ZTK diodes comprise a so-called "adjustment stage" of a kind already described

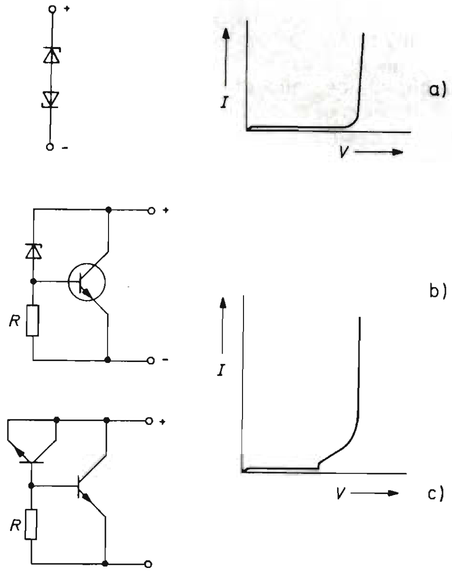


Fig. 36: Methods of temperature compensation
a) Series circuit of several discrete diodes
b) Circuit with Zener diode and shunt transistor, made up of discrete semiconductors
c) Monolithic integrated circuit with Zener diode and shunt transistor

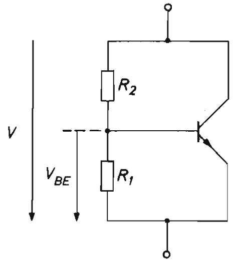


Fig. 37: Adjustment stage

briefly with reference to Fig. 32. Fig. 37 illustrates once more the circuit diagram of such an adjustment stage [27], [28] to which the following expression applies, provided the transistor operates with a sufficiently high current gain.

$$V = V_{BE} \cdot \frac{R_1 + R_2}{R_1} = V_{BE} \cdot \left(1 + \frac{R_2}{R_1}\right) \qquad (34)$$

If the base emitter voltage V_{BE} changes as a result of temperature variations, the voltage V also changes, according to the following expression:

$$\frac{dV}{dT} = \frac{R_1 + R_2}{R_1} \cdot \frac{dV_{BE}}{dT}; \qquad (35)$$

for in a monolithic integrated circuit the ratio of resistors is virtually independent of temperature. With the aid of this adjustment stage, a voltage V is produced which may be any desired – not necessarily integral – multiple of a base emitter voltage V_{BE}, its temperature coefficient being equal to the temperature coefficient of the base emitter voltage of a transistor. This ensures a temperature compensation more accurate than that achievable by means of several diodes in a simple series circuit in which fractions of V_{BE} cannot be realised.

Referring to type ZTK 22, the internal circuit of a ZTK diode (Fig. 38) will now be discussed. Two transistors operating as Zener diodes, each with

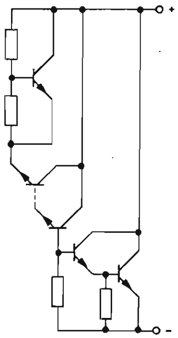
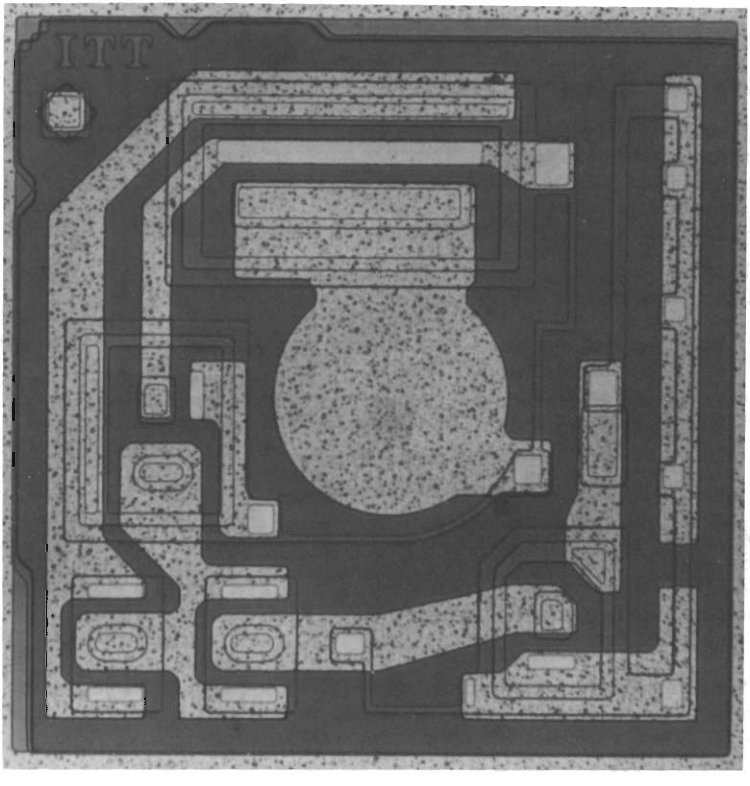


Fig. 38: Internal circuit of the ZTK 22

an emitter to base breakdown voltage of approximately 9 V, are available. To the 18 volts thus obtained must be added a base emitter voltage of the Darlington stage amounting to approximately 1.3 V. This results in 19.3 V, and the residual 2.7 V are taken over by the adjustment stage. For the ZTK 22, a circuit arrangement comprising three transistors acting as Zener diodes would also be feasible, but in this case each of them should have an emitter to base breakdown voltage of 6 V.

The circuit arrangement of Fig. 38 is characterised by a special feature: The collectors of all transistors are at common potential. This does away with the need for electrical separation of the collectors by means of so-called isolation channels. The resulting structure of the ZTK diodes is thus considerably simplified, see Fig. 39. The chip has an area of approximately 0.5 x 0.5 mm², so that it can be easily accommodated in a glass case DO-35 (Fig. 11). A special advantage gained with this small glass encapsulation is the extremely short thermal run-in time of only 20 s, this being an important factor when the stabiliser is employed for the tuning voltage of TV tuners.

Fig. 39: The chip of a ZTK diode, magnification approximately 200 times



2.3.3. Data of the ZTK diodes ZTK 6,8 . . . ZTK 33

The table below contains the most important data of the ZTK series.

Table 4:
Main data of the temperature compensated Zener diodes ZTK 6,8 . . . ZTK 33

Type	Operating voltage at $I_Z = 5$ mA	Dynamic resistance at $I_Z = 5$ mA	maximum operating current [1] at $T_{amb} = 45$ °C
	V_Z V	r_{zi} Ω	I_Z mA
ZTK 6,8	6.4 . . . 7.1	10 (< 25)	36
ZTK 9	9 . . . 10	10 (< 25)	27
ZTK 11	10 . . . 12	10 (< 25)	19
ZTK 18	16 . . . 20	11 (< 25)	13
ZTK 22	20 . . . 24	11 (< 25)	10
ZTK 27	24 . . . 30	12 (< 25)	8
ZTK 33 (≈ TAA 550)	30 . . . 36	12 (< 25)	7

admissible junction temperature	T_j	150	°C
admissible storage temp. range	T_S	– 20 . . . +150	°C
Temperature coefficient of the operating voltage at $I_Z = 5$ mA ± 0.5 mA in the range of $T_{amb} = 20 \ldots 60$ °C	α_{VZ}	– 2 (– 10 . . . +5) [1]	10^{-5}/°C
Thermal run-in time	t_{th}	20 [2]	s
Thermal resistance Junction to ambient air	R_{thA}	< 0.4 [1]	°C/mW

[1]) Valid provided that leads are kept at ambient temperature at a distance of 8 mm from the case.

[2]) At the end of this time ΔV_Z has reached 90 % of its final value ΔV_{Zmax}
$\Delta V_{Zmax} = | V_Z(\alpha) - V_Z(0) |$
where $V_Z(0) = V_Z$ in the instant of turn-on
and $V_Z(\alpha) = V_Z$ at thermal equilibrium.

A comparison with the simple Zener diodes of series ZPD, listed in table 2 on page 32, shows that types ZTK 8 . . . ZTK 33 have an inherent differential resistance which is much lower, and all ZTK diodes a much lower temperature coefficient of the operating voltage. This can also be ascertained from Figs. 40 and 41.

A mathematical comparison of types ZPD 33 and ZTK 33 illustrates the advantage of the ZTK diode. According to equation 6, the ZPD 33 at $I_Z = 5$ mA offers a static differential resistance of $r_{zu} = r_{zi} + r_{zth} = 40\ \Omega + 300\ \Omega = 340\ \Omega$.

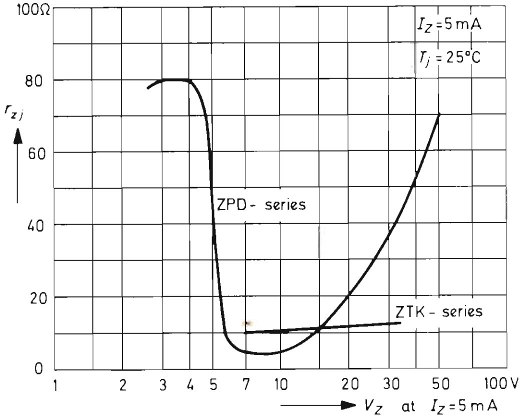


Fig. 40: Inherent dynamic resistance of ZPD and ZTK diodes

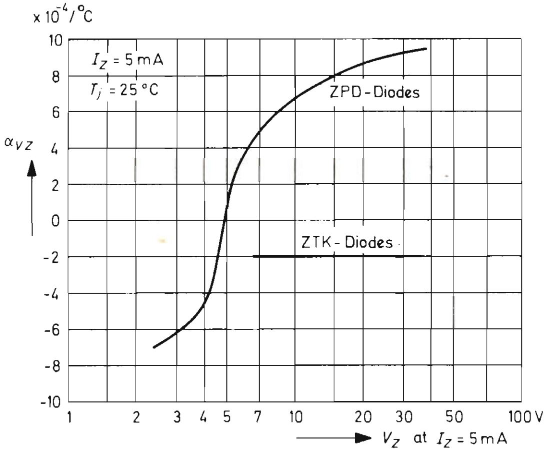


Fig. 41: Temperature coefficient of the operating voltage of ZPD and ZTK diodes

The values for r_{zj} and r_{zth} were taken from table 2 and Fig. 26 respectively. If we assume a supply voltage of 60 V and a load current of 2 mA, then, with $I_Z = 5$ mA, the series resistance in the circuit arrangement of Fig. 24 is

$$R_S = \frac{V_{in} - V_Z}{I_Z + I_{out}} = \frac{27\text{ V}}{7\text{ mA}} = 3.9\text{ k}\Omega$$

and the stabilising factor according to equation 12, is

$$S = \left(1 + \frac{R_S}{r_{zu}}\right) \cdot \frac{V_Z}{V_{in}} = \left(1 + \frac{3900\ \Omega}{340\ \Omega}\right) \cdot \frac{33\text{ V}}{60\text{ V}} = 6.9\,.$$

For the ZTK 33, we obtain from equation 7 the thermal differential resistance

$$r_{zth} = \alpha_{VZ} \cdot R_{thA} \cdot V_Z^2 = -2 \cdot 10^{-5}/°C \cdot 400\ °C/W \cdot 1090\ V^2,$$

and

$$r_{zu} = r_{zi} + r_{zth} = 10\ \Omega - 8.7\ \Omega = 1.3\ \Omega$$

and the stabilising factor is

$$S = \left(1 + \frac{R_S}{r_{zu}}\right)\frac{V_Z}{V_{in}} = \left(1 + \frac{3900\ \Omega}{1.3\ \Omega}\right) \cdot \frac{33\ V}{60\ V} = 1650\,.$$

Thus, S is 240 times higher than in the case of the ZPD 33! In both cases the calculation was based on typical values and no worst-case calculations were carried out.

It only remains to see by what amount the voltage in the circuit arrangement of Fig. 24 under consideration varies as a result of a 25 °C change of temperature. In the case of the ZPD 33, the temperature coefficient of the operating voltage $\alpha_{VZ} = 9 \cdot 10^{-4}/°C$ and the operating voltage varies by

$$\Delta V_Z = \Delta\ T_j \cdot \alpha_{VZ} \cdot V_Z\,, \tag{36}$$

which, in numerical values, reads

$$\Delta V_Z = 25\ °C \cdot 9 \cdot 10^{-4}/°C \cdot 33\ V = 0.74\ V$$

or, relatively, +2.5 %.

In the case of the ZTK we obtain

$$\Delta V_Z = 25\ °C \cdot (-2 \cdot 10^{-5}/°C) \cdot 33\ V = -0.0165\ V$$

or, relatively, − 0.05 %. Even if, according to a worst-case consideration the extreme spread of the temperature coefficient ($-10 \ldots +5 \cdot 10^{-5}/°C$) were taken into account, the least favourable variation of the operating voltage to be anticipated would, in relative terms, be − 0.25 %, i. e. still better by one order of magnitude than in the case of the ZPD 33.

3. Monolithic integrated voltage regulators [14]

3.1. General

The historic development of semiconductor devices used for the stabilisation of direct voltages has followed the logical path from the Zener diode [15] via the monolithic integrated temperature-compensated Zener diode [11], the monolithic integrated voltage regulators with external power transistor to the three terminal monolithic integrated, overload protected voltage regulators now offered with the INTERMETALL series TDD 1605 . . . 1624 for stabilised voltages from 5 V to 24 V.

The advantages of this new series are: A minimum of additional components and compact design. The employment of a transistor case produced in large quantities (Fig. 42) made it possible to realise a favourable price. „On-card regulation" has become possible, i. e. the voltage regulator can now be installed directly in situ – on the individual printed circuit board. This, in turn, brings other advantages, e. g. with regard to power dissipation and in case of failure, because only a single small sub-assembly has to be investigated. Besides, troubleshooting is made easier, and this applies also to automatic troubleshooting processes – a method used more and more frequently in connection with colour television receivers.

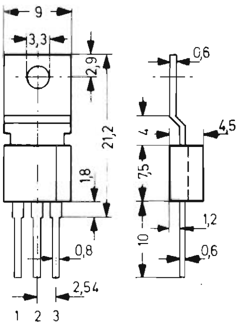


Fig. 42:
Plastic case, similar to 34 A 3
Pin 2 (GND) is connected to the cooling fin.
Weight approx. 1.5 g Dimensions in mm

3.2. Voltage regulators of the series TDD 1605 . . . TDD 1624

3.2.1. Construction

This series of integrated voltage regulators was developed with a view to cater, with one economical device, for the largest number of applications in the medium current range. It comprises regulators for output voltages of 5 V, 6 V, 8.5 V, 10 V, 12 V, 15 V, 18 V and 24 V. The types having output voltages of up to 12 V are designed for output currents I_{op} of at least 500 mA, whereas in the case of types with higher output voltages the output current decreases because of power dissipation, down to 200 mA in the case of the 24 V type. Output voltage tolerance is ± 5 % throughout.

To safeguard a monolithic voltage regulator against destruction or damage from the load side, there are actually four methods which can be employed individually or in combination: 1) simple current limitation, 2) thermal protection, 3) retrograde current limiting and 4) current limitation depending upon the voltage drop. The factor common to all these methods is that the stabilised voltage is recovered after the overload has been eliminated, without any additional external action being necessary.

For the INTERMETALL voltage regulators of the TDD 16 . . . series, overload protection with retrograde current limiting characteristic [31] was chosen. This principle whose characteristic appears in diagrammatic representation in Fig. 43 produces, within the working range, the same behaviour as simple current limitation. However, when overloading occurs it offers the considerable advantage of lower power dissipation. The characteristic is unequivocal, that is to say, within the relevant range, independent of the input voltage. This type of overload protection is mainly used in stabilisers of medium power at currents of up to approximately 0.5 A. It offers the advantage that the internal resistance of the power supply unit is not of critical importance, always provided it is small enough.

To illustrate the power dissipation in the regulator, the power dissipation during normal operation and the power dissipation under short-circuit conditions have been entered in Fig. 43 in the shape of shaded areas. In addition, Fig. 43 shows the hyperbolic power dissipation curve for normal operation. If heat dissipation has been dimensioned for normal operation, only those parts of the characteristic can be exploited which are situated to the left of hyperbola 1. Hence, loading with a resistance of a value between R_{L1} and R_{L2} is prohibited. For $R_{L1} < R_{L2}$, the permitted zone is entered again, and in the case of a full short circuit, power dissipation is even considerably lower than permitted. If total overload protection is desired, the cooling arrangement must be devised according to hyperbola 2. In that case, the whole stabiliser characteristic will be to the left of hyperbola 2, i. e. the power dissipation for all the conceivable operating points is of equal or lower value.

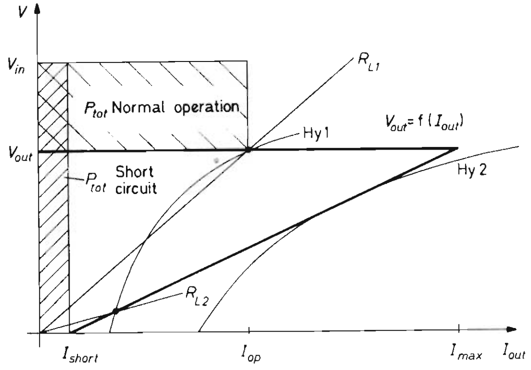


Fig. 43: Characteristic of a voltage regulator with overload protection by retrograde limiting characteristic

In order to make economical manufacture a practical possibility, the same crystal topology is used for all types of the TDD 16... series and the desired output voltage is set by adjusting a resistance matrix by means of the aluminium wiring.

The working of a regulator will now be described with reference to the block diagram of Fig. 44. A temperature-compensated reference voltage for the stabilising circuit is generated by means of a combination of diodes and Zener diodes. The control amplifier compares the desired value (reference voltage) with the actual value (output voltage, or part thereof) and, with its output signal, actuates the correcting element, i. e. the series transistor. Across the measuring resistor R_M, a voltage proportional to the output current is derived which, via an amplifier, effects current limiting with a retrograde characteristic.

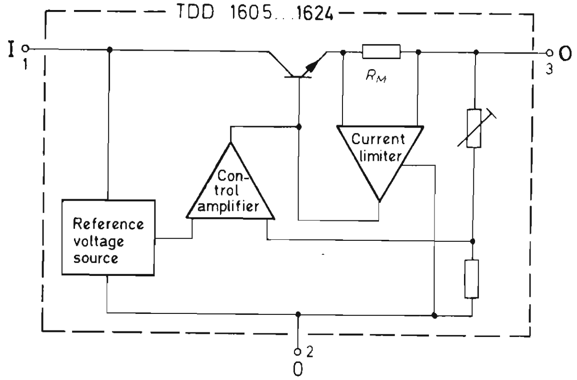


Fig. 44: Block diagram of the integrated voltage regulators TDD 16...

3.2.2. Data for voltage regulators TDD 1605... TDD 1624

Fig. 45 shows the schematic connection diagram, Fig. 46 the schematic characteristic and Fig. 47 the application circuit of integrated voltage regulators TDD 1605... TDD 1624. Their data are given in the table below.

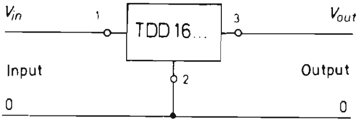


Fig. 45: Schematic connection diagram of the voltage regulators TDD 16...

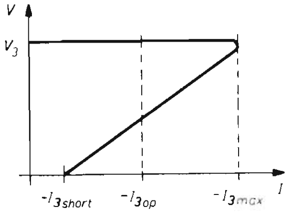


Fig. 46: Schematic characteristic of voltage regulators TDD . . .

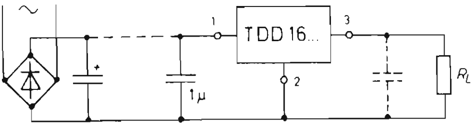


Fig. 47: Application circuit of voltage regulators TDD 16 . . .

Table 5:
Data for voltage regulators TDD 1605 . . . TDD 1624
All voltages are referred to pin 2.

Maximum ratings

Input voltage

TDD 1605 . . . TDD 1618	V_1	35[1])	V
TDD 1624	V_1	40	V
Temperatur of chip	T_j	125	°C

Characteristic at T_C = 25 °C and $V_1 = (V_3 + 5\text{ V})$

Type	Stabilized voltage tol. ± 5 %	Maximum load current	Onset of current limiting	Output resistance
	V_3 V	$-I_{3\,op}$ mA	$-I_{3\,max}$ mA	$\Delta V_3/\Delta I_3$ mΩ
TDD 1605	5	500	1000	75
TDD 1606	6	500	1000	75
TDD 1608	8.5	500	1000	75
TDD 1610	10	500	1000	100
TDD 1612	12	500	1000	100
TDD 1615	15	400	800	150
TDD 1618	18	300	600	150
TDD 1624	24	200	400	200

[1]) Upon request all types are available with an admissible input voltage of 40 V.

Voltage tolerance at $-I_3 = 10$ mA	ΔV_3	± 5	%
Stabilizing factor at $f = 100$ Hz	$\Delta V_1/\Delta V_3$	> 200	
Temperature dependence of the output voltage	$\dfrac{\Delta V_3}{\Delta T_C \cdot V_3}$	10^{-4}	1/°C
No-load current consumption at $I_3 = 0$	I_{1r}	7	mA
Short-circuit current	$-I_{3\,short}$	50	mA
required longitudinal voltage at $-I_{3\,op}$	$V_{1/3}$	> 2.5	V
Thermal resistance			
Chip – cooling fin	R_{thC}	10	°C/W
Chip – ambient air	R_{thA}	70	°C/W

3.3. The car voltage regulator TCA 700 X

3.3.1. Demands placed upon this device

The integrated voltage regulator TCA 700 X was specially designed for stabilised power supplies of motor vehicle instrumentation systems (e. g. fuel or water temperature indication). Since operating conditions in a car are completely different from those obtaining in television receivers or similar equipment, the TCA 700 X had to be developed on completely different lines from the TDD 16.. series. The TCA 700 X, being intended for vehicles with a 12 V electrical supply system, delivers a stabilised output voltage of 10 V. In view of the low voltage drop it is not feasible to isolate the IC from the voltage peaks superimposed on the vehicle's supply voltage by means of an *RC* or an *LC* network. Therefore, the TCA 700 X had to be dimensioned in such a way that it is not damaged by these voltage peaks.

3.3.2. The TCA 700 X – design considerations

The peaks encountered in supplies aboard a vehicle are caused by the interruption of currents flowing through inductances. They have an EMF of several hundred volts and a short-circuit current of a few ampères [32]. One way to safeguard a device to which current is applied from such a source is a power mismatch. In other words, it will be necessary either to provide a device whose blocking voltage is higher than the voltage peaks encountered, or it will be necessary to short-circuit the voltage peaks while keeping the voltage drop as low as possible.

Since reverse voltages of several hundred volts can hardly be realised in bipolar technology while currents of several hundred ampères can be handled quite easily, a decision in favour of shorting the voltage peaks was made when developing the TCA 700 X. In order to accomplish this, a circuit with Zener characteristic is arranged parallel with the input of the actual voltage stabiliser, which circuit has a turn-over voltage of 24 V. In the reverse direction, this protective circuit behaves like a diode with

low bulk resistance, which is capable of absorbing peak currents of up to 15 A (e-function with $\tau = 1$ ms). The voltage regulator and the protective circuit are comprised in one and the same chip, a considerable part of the chip surface being occupied by the protective circuit. The voltage regulator is designed for a load current of 220 mA, operates with an output voltage of 10 V $\pm$0.225 V and is provided with a current limiting arrangement with retrograde characteristic.

Since the integrated Zener diodes cannot be manufactured with a sufficiently close voltage tolerance, the exact output voltage must be set by an individual adjustment process which is carried out in common for all the ICs on one wafer. To this purpose, a finely stepped potential divider is provided on the crystal a part of which is at first short-circuited by the wiring. On each finished wafer containing several hundred ICs, a few spatially distributed ICs are measured according to a predetermined pattern, and the result is evaluated in a computer. The latter determines the appropriate etching process for each wafer, and the correct output voltage on the chip is set during this etching process by means of the aluminium conductors. This method is possible owing to the fact that the voltage tolerances of the Zener diodes on the same wafer differ only very little from one another.

3.3.3. Technical data of the TCA 700 X

The TCA 700 X is contained in a plastic case similar to 34 A 3 (Fig. 42). Its data are listed in the table below. Fig. 48 shows the connection schematic of the TCA 700 X.

Table 6:
Data of the car voltage regulator TCA 700 X

All voltages are referred to pin 2.

Maximum Ratings

Input voltage continuously	V_3	$-0.5\ldots+16$	V
pulsed, max. 1 s	V_3	20	V
pulsed, max. 0.1 ms with $R_i = 100\ \Omega$	V_3	200	V
Input current, pulsed, decaying according to an e-function with $\tau = 1$ ms	$-I_3$	15	A
Junction temperature	T_j	125	°C
Storage temperature range	T_S	$-40\ldots+125$	°C

Recommended Operating Conditions

Load resistance	$R_{1/2}$	> 45.5	Ω

Characteristics at R_{thS} = 20 °C/W [1]), T_{amb} = 25 °C

Stabilized voltage at V_3 = 12 ... 16 V, $R_{1/2}$ = 45.5 ... 330 Ω	V_1	9.775 ... 10.225 V	
at V_3 = 11.5 V, $R_{1/2}$ = 45.5 Ω	V_1	> 9.65	V
at V_3 = 10.8 V, $R_{1/2}$ = 45.5 Ω	V_1	> 8.95	V
Temperature dependence of the stabilized voltage at V_3 = 13.5 V, $R_{1/2}$ = 70 Ω	$\frac{\Delta V_1}{\Delta T_C}$	+1	mV/°C
Current limiting starts at	I_1	> 220	mA
Current consumption at I_1 = 0	I_3	8	mA
Thermal resistance junction to cooling fin	R_{thC}	< 10	°C/W

[1]) R_{thS} is the thermal resistance between cooling fin and ambient air.

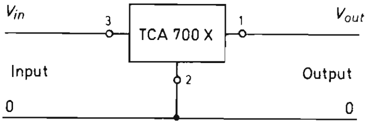


Fig. 48: Connection schematic of the TCA 700 X car voltage regulator

3.4. Cooling of integrated voltage regulators

Fundamentally, the explanations given in sections 2.1.3.2. and 2.1.5.3. apply also in this context. In many cases the dissipation of heat through the IC case alone will not be adequate. The device will have to be screw-connected to a cooling fin or a heat sink, the cooling fin conveying the heat to the ambient air. The minimum size of the cooling fin is calculated as follows, on the basis of the power dissipation encountered in operation. Power dissipation can be calculated from the equation

$$P_{tot} = (V_1 - V_3) \cdot I_L + V_1 \cdot I_{1qu} \quad (37)$$

where
V_1 is the maximum input voltage (for the maximum load current I_L)
V_3 the stabilised voltage
I_L the maximum load current to be encountered
I_{1qu} the quiescent current consumption

The indices 1 and 3 apply to voltage regulators of the TDD 16.. series. For the TCA 700 X, the subscripts 1 and 3 should be interchanged. The required thermal resistance of the cooling sheet or heat sink will then be

$$R_{thS} = \frac{T_j - T_{amb}}{P_{tot}} - R_{thC}. \quad (38)$$

Numerical values for aluminium cooling fins may be gathered from the diagram of Fig. 28. Instead of a cooling fin it would also be possible to provide a heat sink consisting, for example, of an aluminium profile. The construction is considerably simplified if the stabiliser is screw-connected to the thermally conductive chassis. Since the negative terminal of the regulator is connected to the cooling fin, it will be necessary, in the exceptional case of the negative terminal not being connected to the chassis, to mount the regulator with an insulation, e. g. by means of an interposed mica washer. It should be remembered, however, that this will increase the thermal resistance.

4. Application circuits

4.1. Circuit arrangements with Zener diodes, ZTE and ZTK diodes

Since ZTE and ZTK diodes are in the same category as simple Zener diodes as regards their application, the circuit arrangements described here are valid for Zener diodes as well as for ZTE and ZTK diodes. It should be noted, however, that the temperature compensated ZTK diodes are specified for $I_Z = 5$ mA and that they are temperature compensated for this current. Therefore, it is essential for them to operate at this current level.

4.1.1. Parallel stabiliser circuits

Fig. 49 shows the basic circuit for the employment of a Zener diode in a parallel stabiliser circuit. Via a series resistance R_S, the Zener diode is connected to the input voltage V_{in}, and the stabilised output voltage V_{out} derived across the Zener diode. The calculations for this circuit were explained in section 2.1.5. There is no objection to connecting several Zener diodes in series, in order to obtain a higher voltage with tappings, if necessary. By an appropriate choice of Zener diodes, such a series circuit may produce stabilisation of a much higher quality than a single diode having a higher Zener voltage. Example: A ZPD 30 diode has an inherent differential resistance of 35 Ω (table 2) at $I_Z = 5$ mA and a thermal differential resistance of 230 Ω which, together, results in a static differential resistance of 265 Ω. On the other hand, three ZPD 10 diodes, together, have an inherent differential resistance of 15.6 Ω at $I_Z = 5$ mA, and a thermal differential resistance of 60 Ω which, together, results only in about 75 Ω instead of 230 Ω. Here, then, a series circuit produces an effect similar to that described towards the end of section 2.1.5.2. and achieved by choosing a Zener diode of higher power capability.

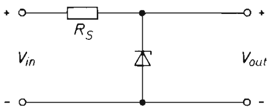


Fig. 49: Parallel stabiliser circuit

If the stabilising factor of the basic circuit (Fig. 49) is inadequate, it is possible, in some cases, to achieve a better result by means of a two-stage circuit (Fig. 50). While the stabilising factor, being the product of the stabilising factors of the several stages, is comparatively high in relation to the fluctuations of the input voltage, it can only be usefully exploited if the junction temperature of the second Zener diode is constant. This necessitates a constant load current and constant ambient temperature, or a constant load current and the employment of a ZTK diode in the second stage.

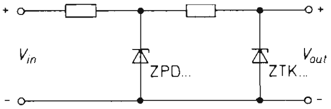


Fig. 50: Two-stage parallel stabiliser circuit

4.1.2. Bridge stabilisers

Considerably better stabilisation factors than those attainable with the basic Zener diode circuit of Fig. 49 can be obtained with bridge circuits (Figs. 51 . . . 53). In fact, if r_z, the differential Zener diode resistance, were constant and the bridge accurately balanced, then the stabilisation factor would be infinite. However, the differential resistance decreases as the diode current is increased. Hence the bridge is usually adjusted in such a way that balance conditions obtain only in the centre of the desired operating range. Under these conditions an increase in input voltage causes the output voltage to increase initially and then, beyond the balance point, to decrease. If the stabilisation factor is measured for a small input voltage swing about a mean input voltage, it will be found that, as the mean input voltage is varied from low to high values, S is initially positive and increases until the balance point is reached when it becomes $+\infty$; S then changes sign and slowly decreases in magnitude from $-\infty$ downwards.

If the stabilisation factor is measured by determining the ratio of maximum input change to maximum output change and the signs are disregarded, then, for the usual ± 10 % input variation, a stabilisation factor of the order of several hundred can be expected. To balance the bridge, an AC voltage should be superimposed on the DC supply and the resistive bridge arms adjusted for minimum AC output. A better balancing method, which also takes the thermal part of the Zener diode differential resistance into account, is to switch the DC bridge input between two levels. In this case, output readings should be taken only after the circuit has reached thermal equilibrium.

In the bridge circuit of Fig. 51, the load is situated in the diagonal of the bridge which consists of a Zener diode and three resistors. The latter should be dimensioned such that, with the bridge operating in the centre of the desired stabilisation range, the condition $R_1/r_z = R_2/R_3$ is met. The differential output resistance of this circuit is approximately $r_z + R_3$; the output voltage is equal to the diode Zener voltage minus the voltage drop across R_3.

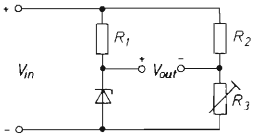


Fig. 51: Bridge stabiliser circuit with one Zener diode

The circuit shown in Fig. 52 is particularly useful as a low voltage stabiliser, the output being approximately equal to the difference of the output voltages of the two Zener diodes and the differential output resistance approximately equal to $r_{z1} + r_{z2} + R_3$. The bridge should be balanced in the centre of the desired stabilisation range, so that $R_1/r_{z1} = R_2/(R_3 + r_{z2})$. Resistor R_3 can be omitted if slight differences in the currents passed by the two bridge arms are acceptable, or if the differential resistance values of the two diodes are approximately equal.

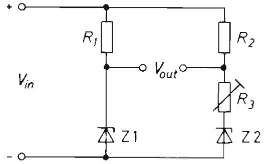


Fig. 52: Bridge stabiliser circuit with two diodes, for a small output voltage

The circuit according to Fig. 53 is recommended when the difference between input and output voltage is only small. The operating voltages of the Zener diodes should be closely matched; the same applies to the values of resistors R_1 and R_2. They should be chosen in such a way that with the bridge operating in the centre of the stabilisation range, they have the same value as the differential resistance of the Zener diodes. The value of resistor R_S should be dimensioned such that sufficient current can flow through the diodes at minimum input voltage, but not so low that at maximum input the Zener diodes are overloaded.

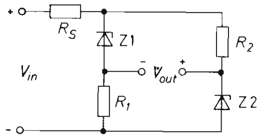


Fig. 53: Bridge stabiliser circuit with two Zener diodes

4.1.3. Voltage stabiliser circuits with Zener diode and transistor

Parallel stabiliser circuits according to Fig. 49 can be designed economically and efficiently only for low loads. Where high loads are involved it is an advantage to "relieve" the Zener diode with a transistor, i. e. the major part of the power dissipated in the stabiliser circuit in the form of heat is generated in a transistor, not in the Zener diode. The degree

of stabilisation that can be achieved by means of the circuits described below is better than in the case of simple parallel stabiliser circuits.

The output power obtainable from the simple Zener diode shunt stabiliser according to Fig. 49 can be increased by applying the stabilised Zener diode output voltage to the base of a power transistor and taking the final output from the emitter of this transistor (Fig. 54). The stabilisation factor of this circuit is determined by the ratio R_1/r_z. Resistor R_2 is merely included to reduce transistor dissipation. The output voltage is equal to the operating voltage of the Zener diode minus the transistor base emitter voltage; the differential output resistance is approximately $(r_z + h_{ie})/h_{FE}$.

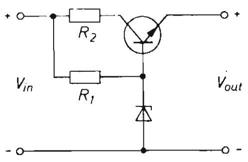


Fig. 54: Simple series stabiliser with transistor and Zener diode

Figs. 55 and 56 show shunt stabilising circuits with Zener diode and power transistor. The output voltage of the circuit arrangement of Fig. 55 equals the sum of the Zener diode operating voltage and the base emitter voltage of the transistor. The resistor in the collector circuit of the transistor may prove useful, in order to reduce the dissipation of the tran-

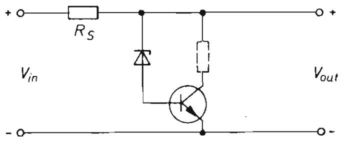


Fig. 55: Shunt stabiliser circuit with Zener diode and transistor, output voltage equals operating voltage of the Zener diode

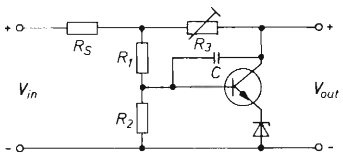


Fig. 56: Shunt stabiliser circuit with Zener diode and transistor, output voltage greater than operating voltage of the Zener diode

sistor. The Zener diode only supplies base current to the transistor. In the circuit arrangement of Fig. 56, the output voltage

$$V_{out} = (V_Z + V_{BE}) \frac{R_1 + R_2}{R_2}, \tag{39}$$

i. e., it is larger than the operating voltage of the Zener diode. The resistor R_3 is of comparatively low value and is preferably trimmed empirically in such a way that the stabilisation factor becomes infinite in the centre of the working range. If R_3 is chosen too high, the output voltage will decline as the input voltage rises. The emitter current of the power transistor passes through the Zener diode. For this reason, usually a power Zener diode will be required. Capacitor C shunts the output of the circuit and, due to the Miller effect, operates with a capacitance increased by the current gain of the transistor.

4.1.4. Current stabilising circuits with Zener diode and transistor

In the stabiliser circuits described so far the intention was to keep the output voltage constant, and, as far as possible, to do so independently of input voltage and load resistance. This required that the source resistance at the output of the stabiliser should be as low as possible. The opposite applies to the current stabilisers described below – the aim being to keep the output current constant, independently of input voltage and load resistance. Consequently, the source resistance at the output ought to be high. Fig. 57 shows the schematic of a constant current source. Its output current is

$$I_{out} = \frac{V_Z - V_{BE}}{R_2} \tag{40}$$

For the stabilisation of the output current in the case of input fluctuations, the ratio R_1/r_{zu} is of prime importance. The source resistance at the output, in general terms, is

$$r_{out} = \frac{R_1 \cdot R_2}{r_{zu}} \tag{41}$$

This simple constant current source forms the basis of a multitude of similar convenient circuit arrangements [4]. Below, a few more current stabilising circuits will be described [9].

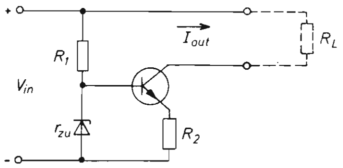


Fig. 57: Schematic of a constant current source

The circuit arrangement of Fig. 58 differs from that of Fig. 57 mainly in that the output current I_{out} can be adjusted by means of potentiometer P1. Potentiometer P2 enables the stabilisation factor dV_{in}/dI_{out} to be adjusted to infinity in the centre of the working range.

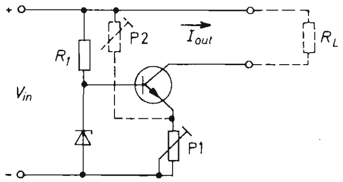


Fig. 58: Adjustable constant current source

If two constant current sources of the type shown in Fig. 57 are connected in parallel, then a two-terminal circuit with current stabilising properties (Fig. 59) is formed which draws a constant current over a wide range of voltages applied to it. Without resistors P and R_2 the two-terminal device would have to be triggered into conduction by the application of a current pulse to one of the bases. This is because the collector current of one transistor is the base current of the other and vice-versa – a condition which must be artificially induced by the application of an external signal. The presence of a resistor R_2 (approx. 1 MΩ) ensures that this happens automatically. Whether R_2 is used or not, the preset control P is important; without it an increase in terminal voltage would result in a slight current rise. This can be compensated by means of potentiometer P. If P is reduced below the optimum value giving maximum current stabilisation, then the two-terminal device exhibits a resistance characteristic which, under certain conditions, is negative. The lower limit of the usable range of the two-terminal constant current circuit is determined by the value $2\,V_Z$, and the upper limit by the value $V_Z \cdot (P/R_1 + 2)$ and by the maximum collector voltage of the transistors.

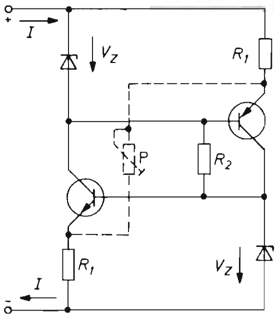


Fig. 59: Constant current two-terminal circuit

4.1.5. Alternating current circuits with Zener diodes

A number of circuit arrangements has been proposed in the case of which one or more Zener diodes limit the amplitude of an alternating voltage. Thus, Zener diodes have been employed in the circuits of Figs. 60 and 61 in connection with a rectifier circuit, in order to obtain from a fluctuating mains AC voltage a constant direct voltage [9].

In Fig. 60, a capacitor is used as a series "dropper" for the Zener diode in connection with a half-wave rectifier circuit. In the forward direction, the Zener diode forms the alternating current path via the input capacitor and, in the reverse direction, it limits the voltage across the charging capacitor. This circuit arrangement is very useful in generating the supply voltage for semiconductor circuits with low current consumption directly from the AC mains. The resistor limits the charging current surge of the series capacitor upon switching-on to a value not detrimental to the diodes. It should be noted that the absence of a line transformer constitutes an electric shock hazard.

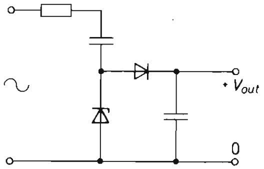


Fig. 60: Stabilising half-wave rectifier circuit

Fig. 61 shows the circuit arrangement of a stabilising rectifier bridge. This circuit can be simplified considerably by designing the mains transformer as a leakage-reactance transformer in which case the required series resistance would be accommodated in the transformer as a power saving reactance. The voltage across the reservoir capacitor is limited to a value equal to the difference between the operating voltage of the Zener diodes and the forward voltage of the conventional diodes. The operating voltages of the two Zener diodes should be identical.

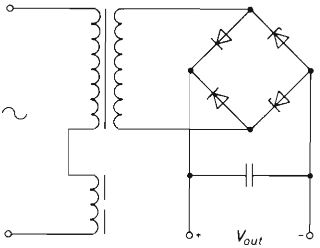


Fig. 61: Stabilising rectifier bridge

In the circuit arrangement of Fig. 62, two Zener diodes with identical operating voltage are series-connected back to back and linked to an alternating voltage source via a series resistance which may also be a capacitive or an inductive reactance. In this way, a symmetrical trapezoidal output voltage is obtained. Its peak value is stabilised. It equals the sum of the forward and the Zener voltage. If the input voltage fluctuates, the steepness of the trapezoidal slopes and, thus, the form factor and the RMS value of the output voltage also vary, however. Therefore, this circuit is only suitable for applications in which the RMS value is of no importance, e. g. as a calibrating source for oscilloscopes.

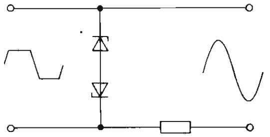


Fig. 62: Symmetrical limiter circuit

Fig. 63 shows an improved variant of the circuit arrangement of Fig. 62. The transformer secondary is provided with a tap (Fig. 63a) and the circuit thus modified to form a bridge. At low input voltages the output then has the near-trapezoidal shape shown in Fig. 63b, the input sine wave being clipped to V_B (the sum of forward and Zener voltage). The compensating voltage appearing at the lower portion of the secondary coil is small. If the input voltage increased considerably, a virtually square output voltage would be obtained in the absence of the compensating coil, with the peak value V_B, its RMS value being considerably higher than in the case of the trapezoidal voltage according to Fig. 63b whose peak value is likewise V_B. However, the compensating coil will now produce the "bucking" voltage indicated by a broken line (Fig. 63c), so that the RMS value of the output voltage (solid line) remains virtually constant, provided the number of turns in the compensating coil is adequate. This circuit is suitable, for example, for stabilising valve filament voltages in precision measuring apparatus.

The simple limiting circuit of Fig. 62 suffers from the disadvantage that in order to obtain good symmetry, the Zener diodes must be quite closely matched. Such diodes are, however, virtually unobtainable in large quantities. Nevertheless, a square or trapezoidal wave form of good symmetry is often required for calibration purposes (e.g. in oscilloscopes). The problem can be solved by shunting a Zener diode across the DC terminals of a bridge rectifier (Fig. 64). When the input goes positive with respect to zero, current passes from terminal 1 via a series resistor R and the diodes D2, D1 and D5 to the zero line, and when it goes negative it passes from zero, via diodes D3, D1, D4 and resistor R, back to terminal 1; the Zener diode consequently always carries a reverse current. This has

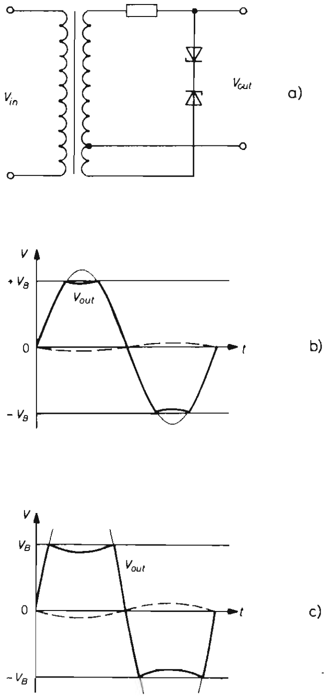


Fig. 63: Stabilising the RMS value of an AC voltage
a) Circuit diagram
b) Output wave form at low input voltages
c) Output wave form at high input voltages

the effect of clipping the output voltage V_{out} to a value determined by the series circuit of the forward voltages of two diodes and the operating voltage of the Zener diode, both in the positive and in the negative direction, i. e.

$$+V_{out} = V_{FD\,2} + V_Z + V_{FD\,5} \qquad (42)$$

and

$$-V_{out} = V_{FD\,3} + V_Z + V_{FD\,4} \qquad (43)$$

By properly choosing operating point and operating voltage for the Zener diode it is possible to attain a certain degree of temperature compensation of the output voltage.

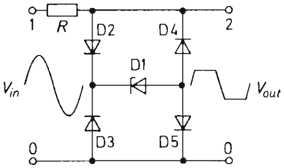


Fig. 64: High symmetry AC stabiliser with only one Zener diode

4.1.6. Limiting and protective circuits

In the circuit arrangements described below, overvoltages which may occur in certain cases are limited by Zener diodes and rendered harmless as far as the protected part of the circuit is concerned. First, let us deal with the limiting of inductive overvoltages which are encountered when interrupting the current passing through inductances. If, in the circuit arrangement shown in Fig. 65, the current in the relay coil is interrupted by the switch, then, in the absence of a Zener diode, an inductive extra-voltage would be produced in the relay winding which, depending on the current interrupted, inductance and capacitance, would amount to several times the battery voltage V_B and would thus endanger the switch and the coil insulation. The Zener diode shunting the relay coil operates with a voltage approximately two to five times the battery voltage and limits the inductive extra-voltage to a value harmless in regard to the contact and the coil insulation; moreover, it produces a controlled speedy drop out of the relay.

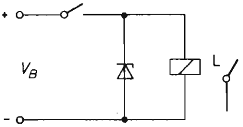


Fig. 65: Zener diode, as a contact protection diode.

In all three variants of the circuit arrangement illustrated in Fig. 66, the transistor is protected against damage caused by the inductive extra-voltage produced on the interruption of collector current. In Fig. 66a, an ordinary silicon diode causes the current to decay in the relay winding and limits the voltage at the collector to the value V_S + approx. 1 V. The dropout time of a relay with such a free-wheeling diode is quite long. In the circuit of Fig. 66b the transistor is protected by a Zener diode whose operating voltage should be higher than the supply voltage V_S, but smaller than the maximum collector voltage of the transistor. The drop out time of the relay decreases with the rising operating voltage of the Zener diode, as related to the supply voltage. This necessitates a suitably voltage-proof transistor. For instance, it would be possible to select $V_{CE0} = 3\ V_S$, and $V_Z = 2\ V_S$, or in order to achieve an even quicker

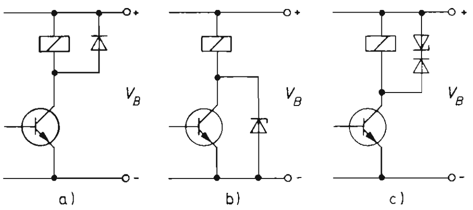


Fig. 66: Protection for a switching transistor with inductive load
a) by means of a free-wheeling diode
b) by means of a Zener diode, connected to the negative terminal
c) by means of a Zener diode, connected to V_S

relay drop out, $V_{CEO} = 4\ V_S$ and $V_Z = 3\ V_S$. The circuit arrangement shown in Fig. 66c has the same effect as that of Fig. 66b, exept that the limiting Zener diode, rather than being connected to the negative terminal, is here connected to the supply voltage V_S. This has the advantage that the power converted into heat in the Zener diode after the blocking of the transistor is smaller in comparison to that of Fig. 66b. So, in some cases, the circuit requires only a small Zener diode. The additional diode is required to prevent the Zener diode from short-circuiting the relay coil when the transistor conducts current.

The circuit arrangement shown in Fig. 66 is suitable, in all three variants, not only for protecting switching transistors driving a relay, but likewise for amplitude clipping and demagnetisation of the primary coil of a pulse transformer, see Fig. 67.

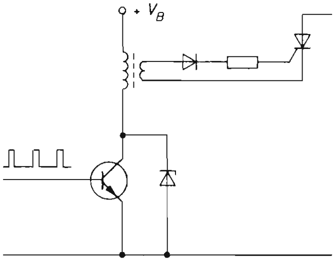


Fig. 67: Amplitude clipping at the primary of a pulse transformer

Whenever an appliance or a circuit component is sensitive to overvoltages on the supply line and must therefore be protected, the employment of one of the following circuits is recommended. In its simplest form, shown in the circuit arrangement of Fig. 68, the operating voltage of the Zener diode is chosen slightly higher than the supply voltage V_S,

by approximately 10 . . . 20 %. This constitutes an effective protection of the load.

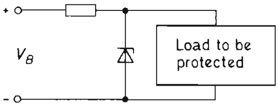


Fig. 68: Protection of a DC load

In cases where the load consumes more power, the circuit arrangements of Figs. 54 and 55 may be used for this purpose. A particularly effective circuit arrangement for major loads is illustrated in Fig. 69. This type of protective circuit, known as a "crowbar" circuit, enforces the separation of the load from mains in that the thyristor fires when the condition of response is reached ($V_S > V_Z$) and short-circuits the supply line, whereupon the fuse responds. The thyristor must be dimensioned for the short-circuit current of the current supply.

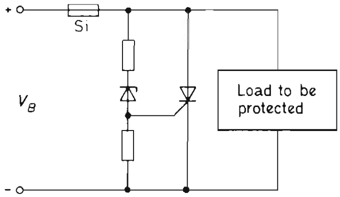


Fig. 69: "Crowbar" circuit

For AC loads, the circuit of Fig. 70 is provided. Two Zener diodes, series-connected with opposite polarity, clip the voltage peaks superimposed on the supply voltage. In Fig. 70a, the protection has been arranged on the primary of the mains transformer and is thus effective for all loads supplied from the secondary coils. In Fig. 70b, only the endangered loads connected to the secondary coils are protected. This may produce the advantage that the leakage inductance of the transformer has a current limiting effect and, therefore, the resistor indicated by broken lines may in some cases be omitted.

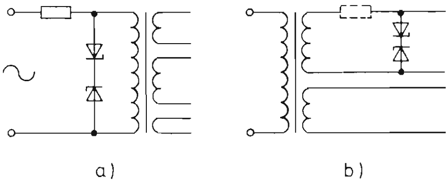


Fig. 70: Protection of an AC load

4.1.7. Various circuit arrangements featuring Zener diodes

In this section, attention will be drawn to several applications of different type which do not fall within the scope of any of the preceding sections.

In the circuit arrangement of Fig. 71, a Zener diode has been employed as a means for overload protection or final range compression in a moving coil instrument. If the resistor R_3 is short-circuited, the current passing through the meter is proportional to the input voltage, as long as the latter does not exceed the Zener voltage. As soon as this is the case, the current through the instrument does not increase further, and the needle remains in its end position. The resistor R_3 causes the final range of the meter to be compressed by an amount which increases with a decrease in the value of R_3. The ratio of the scales on both scale portions will then be $(R_3 + r_z)/(R_1 + R_3 + r_z)$, provided that R_2 or the instrument resistance is large in comparison with all the other resistances.

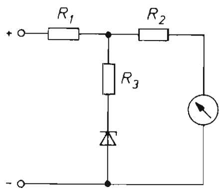


Fig. 71: Zener diode as overload protection or for compressing the final range of measuring instrument

Fig. 72 shows a circuit arrangement for compressing the initial range of a measuring instrument. If resistor R_2 is omitted, current will flow in the measuring instrument only when the input voltage exceeds the Zener voltage, the initial range being completely suppressed. If it is only desired to compress the initial range, then the Zener diode may be shunted by a resistor R_2. For small input voltages the series resistance $R_1 + R_2$ is decisive for the scale dimension, and for large input voltages the resistance $R_1 + r_z$. The parallel circuit of R_2 and r_z may normally be neglected.

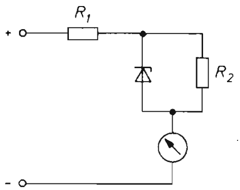


Fig. 72: Initial-range compression for a measuring instrument

In valve amplifiers, the required negative grid bias is normally produced in that a resistor is included in the cathode lead of the valve, said resistor being shunted by a capacitor, so as to avoid negative feedback of the alternating voltage signal. In amplifiers for very low frequencies, this capacitor must be rather large. In DC amplifiers, negative feedback at the cathode resistor cannot be avoided at all. In either case, a Zener diode can be used to advantage for producing the grid bias, as has been shown in Fig. 73.

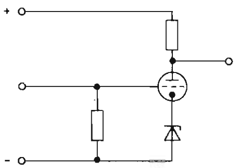


Fig. 73: Producing the grid bias of a valve by means of a Zener diode

If the supply voltage in electronic equipment is to be reduced for a part of the circuit, this can be done by means of Zener diodes. In Fig. 74, for example, the driver transistor which precedes the final transistor has a smaller collector breakdown voltage compared with the final transistor. So, its supply voltage is reduced to tolerable limits by the Zener diode.

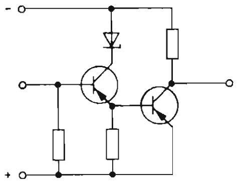


Fig. 74: Zener diode as a series resistance with constant voltage drop

Zener diodes can also be used as coupling elements in amplifiers if there is a potential difference between the output of one stage and the input of the following stage. In Fig. 75 a frequent application of this kind is illustrated. A power transistor must be coupled to a Schmitt trigger. Due to the common emitter resistance of the Schmitt trigger transistors, the collector voltage of the second transistor is higher than the base emitter voltage of the following power transistor, even if the second transistor is fully saturated. A Zener diode, therefore, is ideally suited as a coupling element.

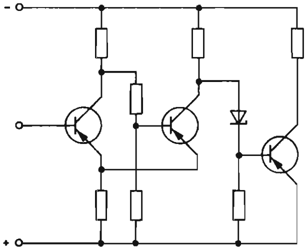


Fig. 75: Zener diode as a coupling element in a switching amplifier

Figs. 76 and 77 show how a Zener diode can be used as a threshold element in voltage monitors. Fig. 76 is an overvoltage monitor system whose lamp will light up when the voltage V exceeds the sum total of Zener voltage, gate trigger voltage of the thyristor and the voltage drop across the series resistor, if only for a brief period:

$$V > V_Z + V_{GT} + I_{GT} \cdot R_S . \qquad (44)$$

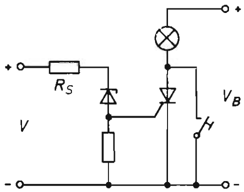


Fig. 76: Overvoltage monitor circuit

When the key is depressed the thyristor blocks and the circuit is again ready to operate. The circuit of Fig. 77, due to the interposed inverting transistor, operates in the opposite sense: The thyristor ignites and the lamp lights up when the voltage V, even briefly, falls below the sum total of Zener voltage, base emitter voltage of the transistor and voltage drop across the series resistor.

Networks made up by several Zener diodes with identical or even with staggered operating voltages may serve to simulate non-linear functions electrically. Thus, by means of the circuit arrangement according to Fig. 78a, the curve $V_{out} = f(V_{in})$ will be approximated by polygonal plotting. A reverse characteristic of such a curve can be obtained by the circuit arrangement of Fig. 79.

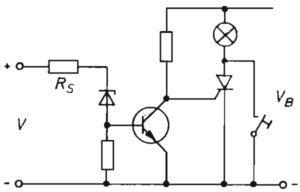


Fig. 77: Undervoltage monitor circuit

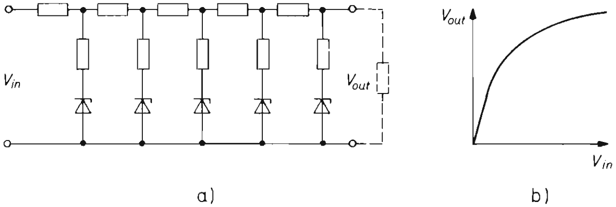


Fig. 78: Function generator
a) Circuit diagram
b) possible characteristic curve

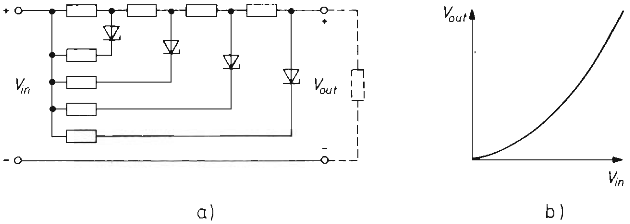


Fig. 79: Function generator
a) Circuit diagram
b) possible characteristic curve

4.2. Voltage regulator circuits

The hitherto described stabilising circuits are designed either for small power operation or merely for low to medium demands on stabilisation factor, generator resistance and temperature coefficient. When dealing with higher powers and higher demands, one of the voltage regulators described below will have to be employed. At first a few basic circuits will be discussed, to be followed by a number of circuit examples.

4.2.1. Basic voltage regulator circuits

The circuits described in this and the following section are elementary circuit arrangements in which the components are not dimensioned and no filter or decoupling capacitors are specified. In practice, it will always be necessary to shunt the output terminals with a small capacitor so as to keep the output impedance for high frequencies low. Moreover, a capacitor for the attenuation of any oscillating tendencies will be required in the control amplifier.

Already in the early days of semiconductor electronics, two basic circuits were developed for electronic voltage regulators which circuits even today serve as the basis for most voltage regulating circuits [33]. Fig. 80 shows the first such circuit in which transistors of the same conductivity type are being used. The reference voltage source is here the Zener diode to which the controlled output voltage is fed and to which the emitter of the control amplifier transistor T2 is connected. A voltage proportional to the output voltage is supplied to the base of T2 via the potential divider R_3/R_4. The final control element, i. e. the series transistor T1, is driven by the collector of the control amplifier transistor. This circuit, in the form described here, is suitable for a fixed output voltage, although this voltage is adjustable within certain limits by varying

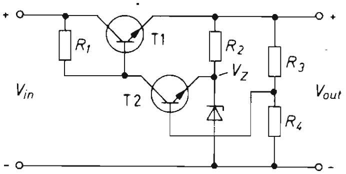


Fig. 80: Voltage regulating circuit with transistors of identical conductivity type

the potential divider R_3/R_4. If, with decreasing input voltage, the difference $V_{in} - V_{out}$ is allowed to shrink considerably, the dimensioning of R_1 becomes problematic in which case the variants described below are recommended. On the other hand, it will always be advisable to keep this difference as small as possible because the heat dissipated in the series transistor T1 is a direct function of this difference. The output voltage V_{out} is approximately

$$V_{out} = V_Z \cdot \frac{R_3 + R_4}{R_4} \quad . \tag{45}$$

The circuit arrangement of Fig. 81 features one NPN and one PNP transistor and works in a manner similar to the above described circuit. Here, the base current for the series transistor T1 is not derived from the input side via resistor R_1, as in Fig. 80, but from the output side via the control amplifier transistor T2, so that the difference $V_{in} - V_{out}$ may assume very

low values, down to the saturation voltage of the series transistor T1. The arrangement is short-circuit proof, for when the output is shorted the control amplifier transistor T2, and therefore also the series transistor T1 are blocked. The output voltage can only be varied within certain limits, in the same way as in the circuit of Fig. 80, by varying the potential divider R_3/R_4, and is approximately

$$V_{out} = V_Z \cdot \frac{R_3 + R_4}{R_3} \, . \qquad (46)$$

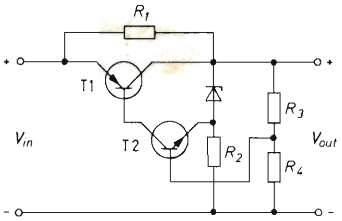


Fig. 81: Voltage regulating circuit with PNP and NPN transistor

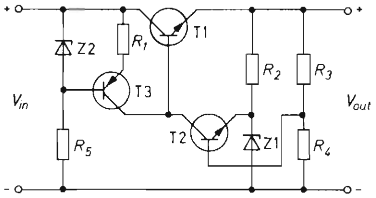


Fig. 82: Improvement to the circuit of Fig. 80, due to the constant current source Z2, T3, R_1

The following circuit details improve the properties of the circuits shown in Figs. 80 and 81 [4]. For example, it may be of advantage to replace the resistor R_1 of Fig. 80 by a constant current source Z2, T3, R_1, as shown in Fig. 82. Even connecting R_1 (Fig. 80) to an auxiliary voltage can improve the circuit properties and facilitate the dimensioning of R_1 (Fig. 83). It may be advisable to stabilise the auxiliary voltage by means of a Zener diode Z2. Furthermore, in the circuit arrangements according to Figs. 80 . . . 83 it is often advantageous to substitute a cascade circuit for the series transistor T1, if the circuit is to be devised for a major load current. The electrical behaviour of the Darlington circuit of Fig. 84a as well as the Lin circuit of Fig. 84b is the same as that of an NPN transistor of high current gain.

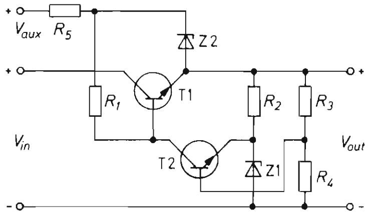


Fig. 83: Improvement to the circuit of Fig. 80 by means of an auxiliary voltage stabilised by Z2

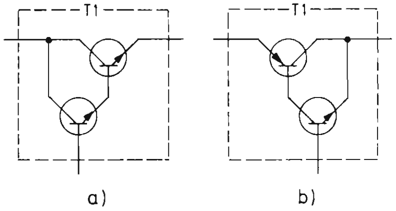


Fig. 84: Cascade arrangement of two transistors, in order to achieve high current gain
a) Darlington circuit (NPN version)
b) Lin circuit (NPN version)

While a series regulator circuit according to Fig. 80 or 81 develops a much better performance than a simple stabilising circuit such as illustrated in Fig. 49 or Fig. 54, its stabilisation factor is only of finite magnitude and the generator impedance at the output is not zero. Two simple steps can bring about a considerable improvement. The stabilisation factor can be considerably increased or even overcompensated so that it goes negative if the circuit of Fig. 80 is expanded by a resistor R_5, see Fig. 85. This resistor enables feedforward control in that a part of the fluctuating input voltage is fed to the input of the control amplifier. The feedforward control system using the resistor R_6 in Fig. 86 applies a voltage proportional to the load current to the input of the control amplifier whereby the generator impedance at the output can be adjusted to zero.

The circuit arrangements of Figs. 80 to 86 feature no particular details which would ensure the independence of the output voltage from temperature. The elements responsible for temperature dependence of the output voltage in these circuits are the Zener diode acting as a reference voltage source and the base emitter diode of the control amplifier transistor T2 if, as can be assumed, the resistors R_3 and R_4 have the same temperature coefficient. The temperature coefficient of the operating voltage of the Zener diode essentially depends on the proper choice of

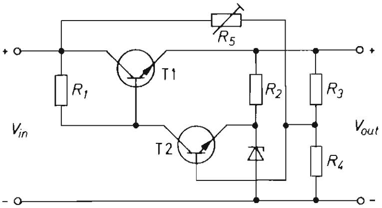


Fig. 85: Improved stabilisation factor, due to feedforward control via R_5

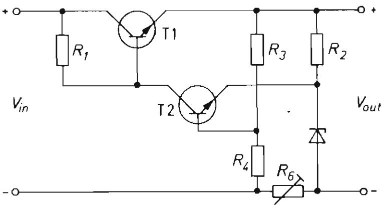


Fig. 86: Reduction of the generator resistance at the output, due to feed-forward control via R_6

this diode, as has already been pointed out in section 2.1.3.1. The base emitter voltage of the transistor depends on temperature in the same way as the forward voltage of a silicon diode, i. e. -2 mV/°C.

In the circuit shown in Fig. 80, the output voltage can be made largely independent of temperature if a 6.2 V Zener diode is chosen, e. g. type ZPD 6.2. From Fig. 17 it can be gathered that the two temperature dependent voltage variations roughly compensate each other. This does not apply to the circuit of Fig. 81, since the operating voltage of the Zener diode and the base emitter voltage of the transistor in Fig. 80 are added to, and in Fig. 81 subtracted from, one another. Therefore, if a certain temperature compensation is desired for the circuit of Fig. 81, a Zener diode with negative temperature coefficient should be used, for example ZPD 3.6 (see Fig. 17).

An independent choice of the voltage of the reference voltage source becomes possible when a temperature-compensated Zener diode is employed and the simple control amplifier T2 (Fig. 80 . . . 86) is replaced by a differential amplifier with two transistors (T2 and T3) such as shown in the basic circuit of Fig. 87. The electrical properties of the differential amplifier can be considerably improved if the (amply dimensioned) re-

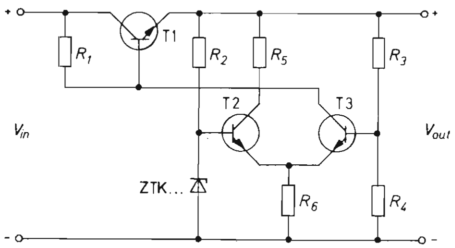


Fig. 87: Improving the circuit of Fig. 80 by a ZTK diode and a symmetrical differential amplifier

sistor R_6 is replaced by a constant current source according to Fig. 57. Alternatively, a resistor of much higher value may be used for R_6 if a negative auxiliary voltage is provided for feeding it (see Fig. 88).

The basic voltage regulator circuits hitherto discussed are designed for fixed output voltages, or output voltages which can only be adjusted over a small range. If the latter is to be adjustable from zero right through to its maximum value, a circuit arrangement similar to that of Fig. 87 is recommended to which must be added the above mentioned negative auxiliary voltage, however. In the circuit arrangement of Fig. 88, the potential drop across potentiometer P, produced by the collector current of transistor T4 represents the reference value. It is continuously variable between zero and full output voltage. Transistor T4 and the Zener diode Z1 form a constant current source according to Fig. 57. The stability of the output voltage V_{out} is a direct function of the output current stability of the constant current source, and it will therefore usually be advisable to pre-stabilise the auxiliary voltage V_{aux}, as has been indicated by a dotted line in the diagram. Moreover, the employment of a Zener diode ZPD 6.2 as diode Z1 gives the smallest temperature dependence of the

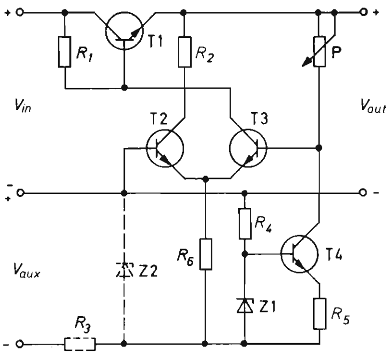


Fig. 88: Voltage regulator with variable output

output voltage. The potentiometer P should be a precision potentiometer with low temperature coefficient. Interchanging the differential amplifier transistors T2 and T3 permits the circuit arrangement according to Fig. 88 to be constructed according to the principle underlying Fig. 81. For the series transistor T1, a PNP transistor will then have to be provided.

4.2.2. Overload protection for voltage regulator circuits

In view of the fact that semiconductor components comprise a very small crystal with extremely small thermal time constant, the series transistor in the voltage regulator circuits described in the previous section are at risk under overload or short-circuit conditions at the output. Ordinary fuses usually react only when the transistor has already been damaged, and it is therefore advisable at any rate to provide electronic overload protection. By using a resistor in the negative lead it becomes possible to expand the circuit arrangement of Fig. 80 into a short-circuit proof arrangement. The first variant is shown in Fig. 89. This circuit comprises, in addition, transistor T3 which remains blocked during normal operation. If, due to an excessive load current, the potential drop across R_6 rises to about 0.6 V, T3 starts to conduct and thus reduces the base current of T1. The arrangement acts as a current limiter, as has already been discussed in section 3.2.1. Under short-circuit or overload conditions, the series transistor is subject to high thermal stress and will have to be dimensioned accordingly. After the removal of the overload or short-circuit condition, the output voltage is restored automatically.

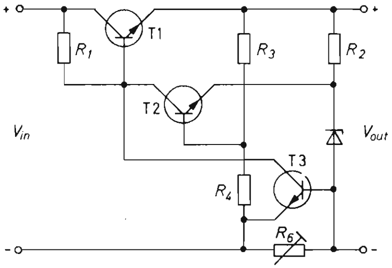


Fig. 89: Circuit arrangement of Fig. 80, with additional output current limiter

As far as the thermal dimensioning of the series transistor is concerned, it would be convenient to block this transistor entirely if an overload or a short-circuit condition occurs, so that an output current would not flow any longer. This is made possible by the circuit arrangement illustrated in Fig. 90. The latter is based on the elementary circuit of Fig. 81 which has been expanded by the automatic turn off consisting of transistors T3 . . . T5. When the potential drop across resistor R_9, which is proportional to the load current, exceeds the base emitter threshold of transistor T3 (about 0.6 V), collector current begins to flow in this transistor, and

thus base current in transistor T4. The collector current of T4 increases the base current of T3. Due to this feedback, the bistable trigger consisting of transistors T3 and T4 is rendered conductive, whereby the auxiliary transistor T5 is rendered conductive at the same time and the base of T2 approaches zero potential. As a result, transistor T2 of the control amplifier and series transistor T1 are blocked, and an output current does not flow any longer. By disconnecting the supply voltage V_{in} the bistable trigger T3, T4 returns to the blocked condition, and the circuit is again ready to operate.

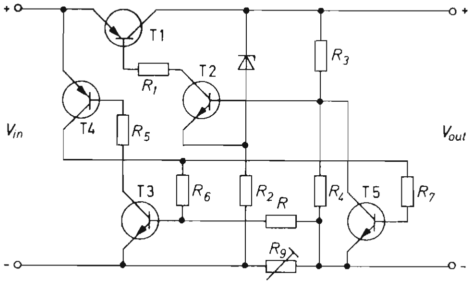


Fig. 90: The circuit arrangement of Fig. 81, with an additional overcurrent cutoff

A circuit arrangement whose working principle can be gathered from Fig. 91 operates in a similar way [31]. In contrast to the previously described circuit of Fig. 90 it may prove advantageous that, after eliminating a short-circuit at the output, the supply voltage need not be turned off because this circuit arrangement will return to the normal operating condition automatically. Fig. 92 shows the characteristic of this circuit. If the output current $I_{out\,max}$ is exceeded as a result of too small a load resistance, the output current jumps to I_{short} and the output voltage assumes a value between zero and V_{short} which depends on the given load

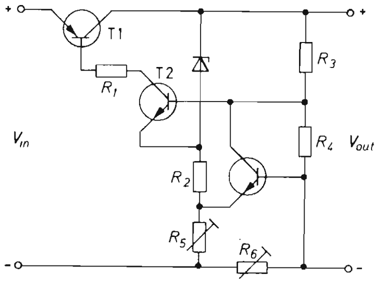


Fig. 91: Circuit according to Fig. 81, with additional built-in overcurrent cutoff, retrograde current characteristic and automatic voltage restoration

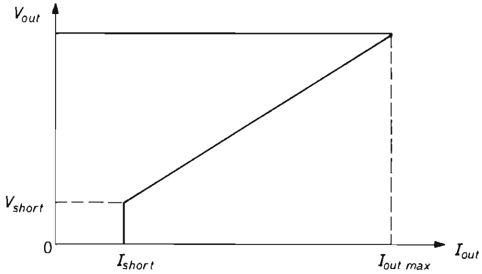


Fig. 92: Schematic characteristic according to the circuit of Fig. 91

resistance. If the load resistance rises, so that the product $R_L \cdot I_{short}$ exceeds the value V_{short}, the output voltage jumps back to its nominal value.

With reference to Fig. 93, a further version of overload protection will now be shown. Here, a planar thyristor is used as the cutoff element [34]. When the potential drop across the resistor R_6 exceeds the ignition voltage of the thyristor (approx. 0.7 V) due to overloading, the thyristor fires and thus grounds the base of the series transistor T1, so that the latter is blocked. A brief interruption of the supply voltage V_{in} renders the circuit operational again, after the overload has been removed.

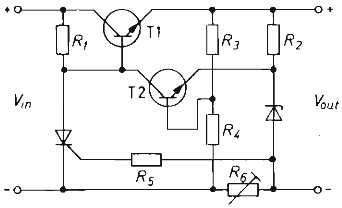


Fig. 93: Circuit of Fig. 80 with additional built-in overcurrent cutoff by means of a thyristor

4.2.3. Circuit examples for voltage regulators

Whereas in the two preceding sections elementary circuits were presented which, when applied in practice, will have to be dimensioned according to given requirements, we shall now present a number of dimensioned circuits, well proven in practical operation [9].

Fig. 94 shows a simple series stabilising circuit according to Fig. 54 in which a Lin circuit (see Fig. 84b) has been employed as series transistor. For an input voltage range of 15 V ± 20 %, the output voltage is ap-

proximately 6 V, and a maximum output current of 1 A is admissible. Measurements of the source impedance at the output revealed:

2 Ω in the range I_{out} = 0 . . . 50 mA

0.1 Ω in the range I_{out} = 50 . . . 500 mA

0.02 Ω in the range I_{out} = 0.5 . . . 1 A

The stabilisation factor S at I_{out} = 0.5 A is greater than 50. Transistor T2 requires a heat sink or a cooling sheet with a maximum thermal resistance of 15 °C/W.

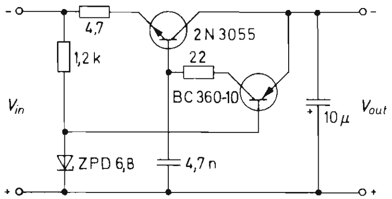


Fig. 94: Simple series stabilising circuit for 6 V, 1 A

The voltage regulator of Fig. 95 is designed for an input voltage range of 17 . . . 22 V and supplies an output voltage of 12 V for a maximum load current of 2 A. It corresponds to the elementary circuits of Figs. 85 and 86, i. e. it comprises two feedforward circuits. By means of the 100 kΩ potentiometer the stabilisation factor is adjusted to its maximum value. By means of the variable 0.15 Ω resistor the source impedance at the outpout can be set to a minimum and the output voltage set to a nominal value by means of the 250 Ω potentiometer. The series transistor 2 N 3055 requires cooling means having a thermal resistance of not more than 5 °C/W and a star-shaped cooling element must be fitted to the BC 140. Maximum ambient temperature is 45 °C at full load.

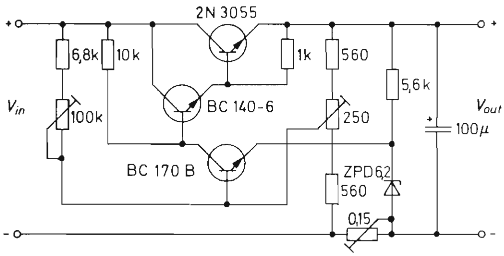


Fig. 95: Voltage regulator circuit for 12 V, 2 A

Fig. 96 shows a short-circuit proof voltage regulator based on the principle of the diagram shown in Fig. 81, designed for an output voltage of 24 V. Input voltages of − 27 . . . − 36 V are admissible. The maximum output current at which the output voltage breaks down may be roughly calculated by using the following equation:

$$I_{out\,max} = \frac{h_{FE} \cdot (V_{out} - V_Z)}{R} \tag{47}$$

wherein h_{FE} is the current gain of the power transistor, taking into account the 3.9 kΩ base-emitter resistance. By means of the 250 Ω potentiometer the output voltage is adjusted, and by means of the 1 kΩ potentiometer the cutoff current. The latter is of the order of 1 A.

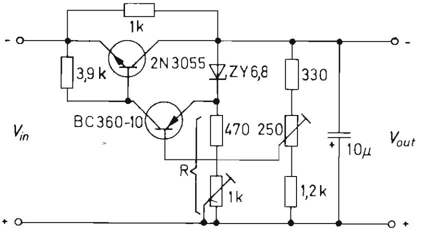


Fig. 96: A short-circuit proof voltage regulator for 24 V, I_{out} = approx. 1 A

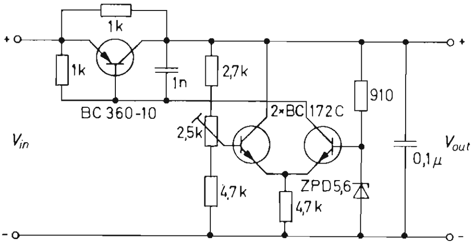


Fig. 97: Voltage regulator with small residual voltage, for 9 V, 50 mA

The voltage regulator circuit of Fig. 97 likewise resembles the elementary circuit of Fig. 81, but is equipped with a differential amplifier similar to the circuit arrangement of Fig. 87. The most important feature of this circuit is that it is conceived fo a very small difference between the input voltage V_{in} and the output voltage V_{out}. The input voltage range is 10 . . . 14 V, and the output voltage 9 V, with an admissible load current of 50 mA.

In the circuit arrangement of Fig. 98, the output voltage can be set to within 6 and 15 V, and the admissible load current is 0.6 A. The input

voltage may fluctuate between 20 and 22 V. An overload cutoff circuit according to the method illustrated in Fig. 93 is provided. Thyristor BRX 44 ignites when the voltage drop across a 1 Ω resistor exceeds the value of about 1 V and will then ground the base of the Darlington series transistor, so that the latter is blocked. This condition is preserved until the thyristor is extinguished by an interruption of the supply voltage V_{in}. The 1 Ω resistor in the negative lead, moreover, effects feedforward control according to the principle illustrated in Fig. 86, and this produces a very low source impedance at the output.

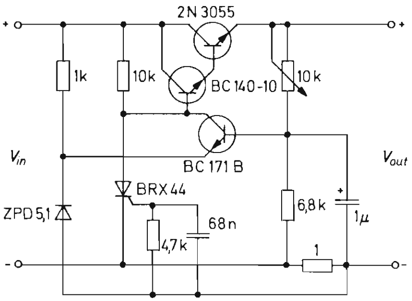


Fig. 98: Voltage regulator with electronic overcurrent cutoff for 6 . . . 15 V, 0.6 A

A critical design problem arises if a voltage stabilising circuit is expected to produce a stabilised voltage of 1.15 V when it is fed from a monocell which, during operation, decays from a maximum of 1.8 V down to 1.2 V. Demands of this kind are made in connection with electronic watches. INTERMETALL has developed the circuit shown in Fig. 99 for this kind of application. Here, the reference voltage source is the base emitter path of transistor T2 which, via transistor T1, controls the series transistor T3. Taking into account the product of the current gain of transistors T1 and T3, the 1 MΩ base resistor of transistor T1 is dimensioned such that the series transistor T3 is fully saturated when transistor T2 does not conduct current. When the output voltage has reached its desired value, a

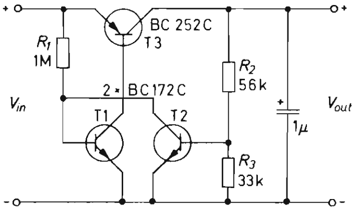


Fig. 99: Voltage regulator for a low battery voltage, output voltage 1.15 V, output current 5 mA

collector current flows in T2 which reduces the base current of transistor T1 to such an extent that the output voltage remains constant at 1.15 V. The reference voltage is the base emitter voltage of transistor T2 amounting to about 420 mV for a collector current of about 1 μA. According to the potential divider ratio R_2/R_3, an output of 1.15 V is achieved. This voltage is kept stable for load currents up to about 5 mA. If resistor R_1 is reduced, higher load currents become admissible. The source resistance at the output is about 1 . . . 2 Ω. A variation of input voltage in the region of 1.2 . . . 1.8 V causes the output voltage to vary by about 70 mV.

Conventional series regulators are unable to cope with an appreciable back EMF for reverse current produced by the load. If the emitter of the series transistor used in the regulator is directly connected to the output (see Fig. 80), then this transistor is likely to be damaged whenever the back EMF exceeds the output voltage by more than the maximum permissible emitter base voltage of the series transistor (usually 5 . . . 7 V). The internal resistance of the supply source to reverse currents is very high. Shunt stabilisers, on the other hand, whilst not suffering from the same disadvantage, pass a continuous quiescent shunt current of the same order of magnitude as the output current. The combined circuit shown in Fig. 100 largely overcomes both these disadvantages. In principle it is very similar to a transformerless output stage familiar from audio engineering, and consists, in the main, of a Darlington pair formed by transistors T1 and T2. A stabilised voltage derived from Zener diode ZPD 5.6 is applied to the base of T1. The series-connected stabiliser diode ZTE 1.5 compensates the base emitter voltages of T1 and T2, so that the output voltages V_{out} is roughly equal to the voltage across the Zener diode and has a low temperature coefficient. Furthermore, a PNP parallel transistor T3 is provided. The common bias of the complementary transistors T2 and T3 is determined by the voltage drop across diode BA 170 and the 100 Ω potentiometer. The quiescent current through transistors T2 and T3 is set to about 10 mA by means of the potentiometer. Under normal conditions the circuit functions like any conventional series regulator and passes a quiescent current of 10 mA. If, however, a back EMF is applied to the output, then the emitter potentional of transistor T3

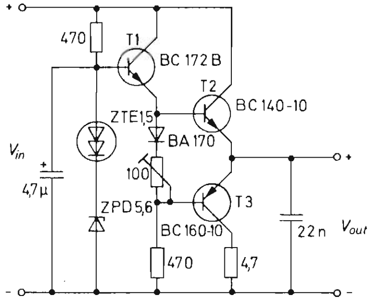


Fig. 100: Combined series and shunt regulator circuit for 5.6 V, ± 200 mA

goes positive, its conductivity rises, and a current passes, via T3, to the zero terminal. Each of the output transistors T2 and T3 should be fitted with a heat dissipator. The range of input voltages is 12 V ± 20 %, and the output voltage is 5.6 V. Load currents of up to 200 mA at the output are admissible in both directions.

The next circuit diagram (see Fig. 101) illustrates a voltage regulator similar to that of Fig. 88, its output voltage being adjustable between zero and 15 V. The admissible output current is 1.5 A and the rated input voltage between 17 and 20 V. The negative auxiliary voltage should be between 16 and 20 V. A Lin circuit similar to that shown in Fig. 84 ist incorporated to act as series transistor and behaves like a high gain PNP transistor. For an ambient temperature of 45 °C the 2 N 3055 requires a cooling sheet having a thermal resistance of no more than 3.5 °C/W and a star-shaped heat dissipator must be fitted to the BC 360. The stabilisation factor of the circuit is approximately 100, and the source resistance at the output approximately 50 mΩ.

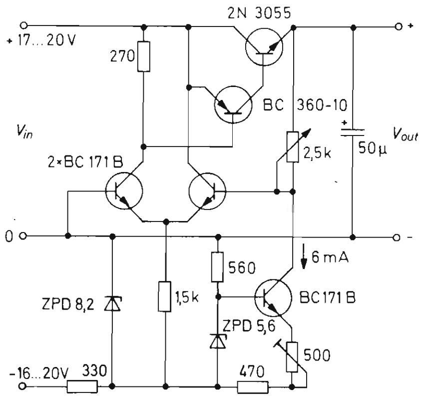


Fig. 101: Voltage regulator circuit for 0 . . . 15 V, 1.5 A

4.2.4. Circuit examples for voltage regulators with integrated circuits

The monolithic integrated voltage regulators developed by INTERMETALL have already been discussed in section 3. Below, a few application circuits will be described. The standard application circuit has already been illustrated in Fig. 47. The capacitor at the output may be omitted in most cases, unless a small output resistance is required, especially at high frequencies. If the charging capacitor of the power supply unit is physically separated from the voltage regulator, an additional capacitor of about 1 µF, directly at the regulator input, will be needed.

Additional external circuitry enables voltage regulating circuits to be constructed with the ICs of series TDD 1605 . . . TDD 1624 which deliver

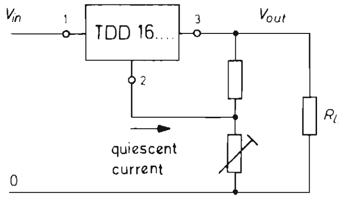


Fig. 102: TDD 16.., used as a voltage regulator with adjustable output voltage

a higher voltage than that for which the IC is rated. Fig. 102 shows the simplest solution, although in this case the quiescent current from terminal 2 of the integrated voltage regulator flows into the potential divider and influences the output voltage. The expanded circuit of Fig. 103 does not suffer from this disadvantage.

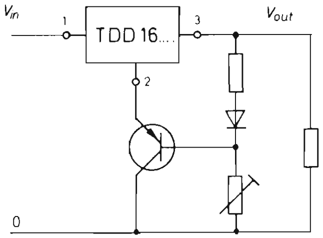


Fig. 103: The circuit of Fig. 102, its function improved

In the circuit of Fig. 104 the integrated voltage regulator is used in the construction of a constant current source. The output current I_{out} is calculated from the equation

$$I_{out} = \frac{V_{3/2}}{R} + I_0 \qquad (48)$$

and, if a variable resistor or a potentiometer is used for R, is continuously variable.

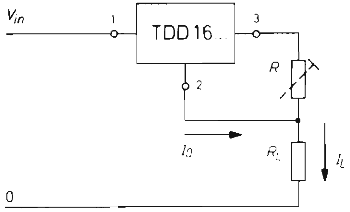


Fig. 104: Constant current source with TDD 16..

4.3. Laboratory power supply unit with voltage and current regulation

Shortly before this book was published, instructions for building a laboratory power supply unit were published [35] which, over the range of 0 . . . 30 V, operates with a continuously variable output voltage and a maximum load current of 2 A. This circuit arrangement incorporates overload protection and permits selection of two operating modes, i. e. cutoff upon overloading or current limiting. The current limiter is designed in such a way that within the range of current stabilisation, the output current is kept constant in the same degree as the output voltage in constant voltage operation.

The circuit (Fig. 105) contains three light-emitting diodes (LED) which indicate the mode of operation. The switch S2 enables changeover between the operating modes "current limiting" (position 1) and "cutoff" (position 2). In position 1, the circuit operates below the current set by means of P3, acting as a constant voltage source (the green LED lights up), and above the set current the red LED lights up to indicate constant current operation. When switch S2 is set to 2, then, during normal operation, none of the LEDs will light up: It is only after a cutoff due to overloading that the yellow LED lights up. If switch S2 is changed over to position 1, the power supply is rendered operative again, after disconnection due to overloading.

Coarse and fine adjustment of the output voltage is possible by means of potentiometers P1 and P2 respectively. In the constant current mode, coarse adjustment of the output current is possible by means of switch S1 in the ranges 0 . . . 0.2 A and 0 . . . 2 A, whereas it is continuously variable by means of potentiometer P3. The series transistor is made up by two power transistors 2 N 3055, connected in parallel, each of them being mounted on a heat sink having a thermal resistance of not more than 2 °C/W. If an output current of 1 A is sufficient, one of the two series transistors may be omitted. The driver is also a 2 N 3055 because this type of device is robust and less expensive than similar, possibly smaller, power transistors. It may be mounted on the heat sink of one of the series transistors.

The current range is set by means of the 2.5 kΩ trimming potentiometer, and the voltage range by means of the 1 kΩ trimming potentiometer. In order to measure the output voltage, a voltmeter with 30 V f. s. d. may be shunted across the output terminals. In order to measure the current, a moving coil instrument may be connected in parallel with the shunt of 4.7 Ω, its full scale deflection being set by means of a series resistance to 1.02 volts.

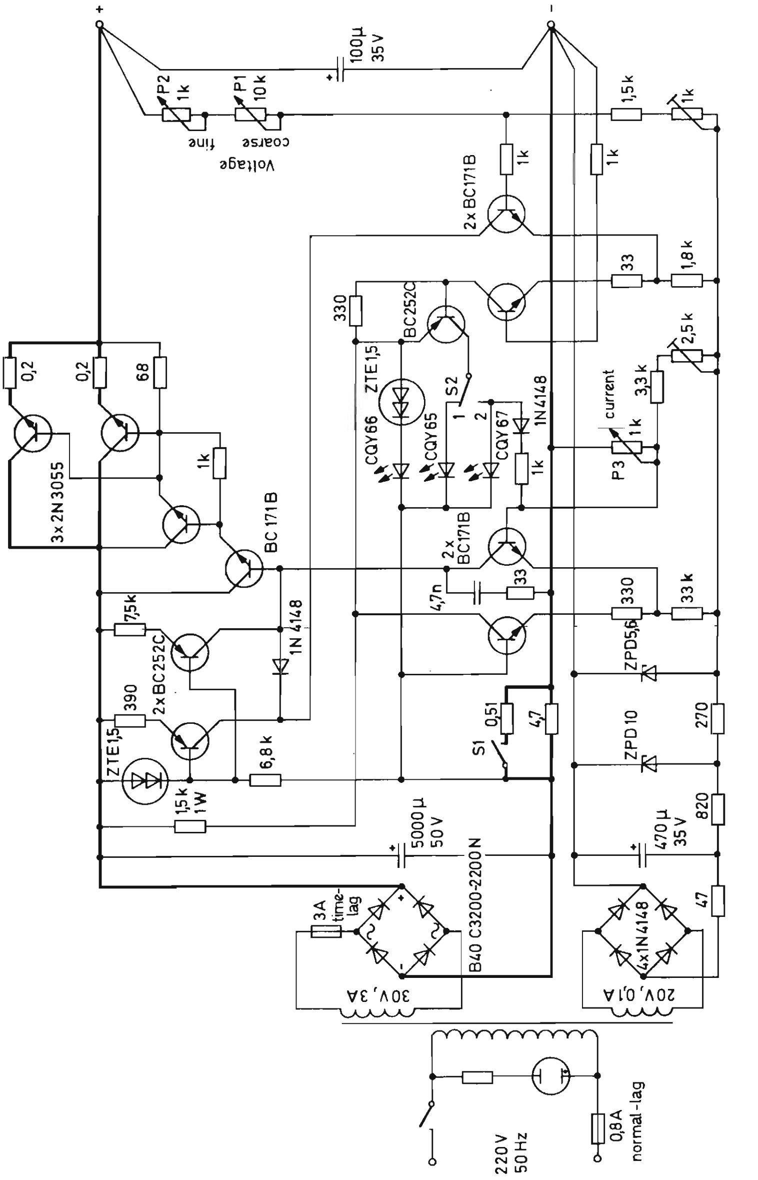


Fig. 105: Laboratory power supply unit, with 0 . . . 30 V output voltage and up to 2 A output current

5. Bibliography

[1] DIN 41 855, Semiconductor devices and integrated circuits, kind of devices and general terms (German standards)

[2] Semiconductor catalogue, INTERMETALL Gesellschaft für Metallurgie und Elektronik mbH, Düsseldorf (Germany), Edition April 1957.

[3] Gerlach, A.: Die Zenerdiode (The Zener diode). ionen + elektronen 1960, No. 9, p. 12 to 15.

[4] Wagner, S. W., Stromversorgung elektronischer Schaltungen und Geräte (Power supply for electronic circuits and equipments). R. v. Decker's Verlag G. Schenck, Hamburg 1964. Mentions further literature.

[5] Funktechnische Arbeitsblätter Wk 22, Franzis Munich (Germany).

[6] Funktechnische Arbeitsblätter Re 11, Franzis Munich (Germany).

[7] Streng, K.: abc von Elektronenröhre und Halbleiterbauelement (abc of the electronic tube and of the semiconductor device). Militärverlag der DDR, Berlin 1973.

[8] Hahn, H.: Thermistoren (Thermistors). R. v. Decker's Verlag G. Schenck, Hamburg und Berlin 1965.

[9] –, –: Discrete Semiconductor Circuit Examples. ITT INTERMETALL Book, 1973, Order No. 6200-99-1E.

[10] Dolega, Dr. U.: Semiconductor Diodes. Funkschau 1974, No. 20, p. 789 to 791, No. 21, p. 819 and 820, No. 22, p. 857 and 858 or ITT Technical Information Semiconductors, Order No. 6200-15-1E.

[11] Dolega, Dr. U.: Temperature-compensated Z-diodes. ITT Technical Information Semiconductors, Order No. 6200-73-1E.

[12] Schilling, W.: Der Transduktor (The transductor). R. Oldenbourg, Munich 1958.

[13] –, –: Thyristoren, Triacs, Triggerdioden, Grundlagen und Anwendungen (Thyristors, triacs, trigger diodes, basics and applications). ITT INTERMETALL Book, 1976, Order No. 6210-01-1D.

[14] Höhn, W.: Monolithic voltage stabilisers. Funkschau 1976, No. 8, p. 293 to 296 or ITT Technical Information Semiconductors, Order Nr. 6200-138-1E.

[15] Gerlach, Dr. A.: Eigenschaften und Anwendungsmöglichkeiten der Z-Diode (Properties and applications of the Zener diode). Bull. S. E. V. 1962, No. 25, p. 1228 to 1237 or ITT Technical Information Semiconductors, Order No. 6200-14-1D.

[16] Zener, C.: A theory of electrical breakdown of solid dielectrics. Proc. Roy. Soc. (London) series A 145 (1934) 855, p. 523 to 529.

[17] McKay, K. G.: Avalanche breakdown in silicon. Phys. Rev. 94 (1954) 4, p. 877 to 884.

[18] –, –: Capacitance diodes, Tuner diodes, Diode switches, PIN diodes, Basics and Applications. ITT INTERMETALL Book, 1976, Order No. 6220-09-1E.

[19] Dietrich, B.; Lehmann, M.: Silizium-Epitaxie-Planar-Transistoren-Technologie und Eigenschaften (Silicon epitaxial planar transistors – technologie and properties). radio mentor 1963, No. 10.

[20] Keller, H.; Micic, L.; Wieczorek, G.: Halbleiterbauelemente zum Erzeugen konstanter Vergleichsspannungen (Silicon devices for producing constant reference voltages). ETZ Ausgabe B, 14 (1962) No. 12, p. 327 to 332.

[21] Höhn, W.: Monolithic integrated voltage stabilisers. Funkschau 1976, No. 8, p. 293 to 296 or ITT Technical Information Semiconductors, Order No. 6200-138-1E.

[22] Stöhr, H. J., Bemerkungen zum Stabilisierungsverhalten von Zenerdioden (Notes on the stabilising characteristics of Zener diodes). Elektronische Rundschau 1962, No. 7, p. 297 to 301.

[23] ITT INTERMETALL manual "Diodes, Zener Diodes, Rectifiers, Thyristors" Order No. 6220-10-1E.

[24] Funktechnische Arbeitsblätter HI 22, Franzis Munich (Germany).

[25] Schach, R.: Thermische Widerstände von Transistoren und Kühlkörpern (Thermal resistances of transistors and heat sinks). Funkschau 1967, No. 5, p. 137 to 140.

[26] Haas, G.: Kühlflächen für Leistungshalbleiter (Cooling surfaces for power semiconductors). Elektronik 1966, No. 5, p. 139 to 141.

[27] Bergtold, F.: rtp-Elektronik-Lehrgang. Regelungstechnische Praxis und Prozeß-Rechentechnik (rtp electronic course. Practical control technique and process computing) 1970, No. 6, p. M 45 to M 48.

[28] –, –: Grundsätzliche Schaltungskonzepte monolithisch integrierter Linearschaltungen (Basic concepts of monolithic integrated linear circuits). Funk-Technik 30 (1975) Nr. 11, p. 308 to 311.

[29] Funktechnische Arbeitsblätter Ba 21, Franzis Munich (Germany).

[30] Uhlmann, S.: Spannungsstabilisierung mit integrierten Schaltungen (Voltage stabilisation with integrated circuits). Bauelemente der Elektronik 3 (1968) No. 15, p. 12 and 13.

[31] Mielke, H.: Überlastschutz für transistorstabilisiertes Netzgerät (Overload protection for a transistor-stabilised power supply unit). Radio-Fernseh-Phono-Praxis 1966, No. 6, p. 189 to 191 or ITT Technical Information Semiconductors, Order No. 6200-33-1D.

[32] Höhn, W.: SAY 115 – An integrated circuit replaces the tacho shaft. Funkschau 1973, No. 9, p. 301 to 304 or ITT Technical Information Semiconductors, Order No. 6200-114-1E.

[33] Hunter, L. P.: Handbook of Semiconductor Electronics. McGraw-Hill, New York, Toronto, London 1956.

[34] Peltz, G.: Circuit examples comprising silicon planar thyristors. Elektronik 1975, No. 10, p. 91 to 95.

[35] Flamm, P.; Reimer, C.: Universalnetzteil mit Stromregelung (Universal power supply unit with current stabilisation). Funkschau 1976, No. 23, p. 109 to 112. A suitable P. C. board (designation FS 4) will be supplied by Funkschau (Munich, Germany) against payment. The publication also contains a detailed part list.

Semiconductor
worldwide
ITT